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THE F.M. SECTION OF MODERN BROADCAST RECEIVERS

I. CIRCUITRY AND STRUCTURAL DESIGN

by H. de QUANT and P. ZIJP.

II. THE BUILT-IN F.M. AERIAL

by M. HUISSOON.

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In the early days of broadcasting the thing which perhaps appealed most to the imagination was the idea of receiving transmissions direct from distant countries. Nowadays the attitude is less romantic: enquiries show that the vast majority of present-day listeners tune in to only a few stations (mainly in their own country) where they can be assured of good reception. The emphasis now lies on reception of good quality. Efforts in this direction are almost all concerned with the use of frequency modulation (F.M.) for broadcasting. Although this system implies short wavelengths (metre waves), whose range is limited, it has important advantages: the signal-to-noise ratio is much better, and, in comparison to most long and medium wave transmitters, the audio spectrum is much wider. Nearly all European countries have a short-wave F.M. service in operation in addition to the usual amplitude modulated transmissions on long, medium and short waves. Accordingly, most of the latest receivers in Europe (with the exception of those in the lowest price class) are adapted for both systems.

It is usually assumed that broadcasting with frequency modulation (F.M.) owes its growing popularity to the fact that with this modulation system the audio spectrum extends to about 15000 c/s, which is much higher than that of the majority of broadcasting transmitters working with amplitude modulation (A.M.). Up to a point this is true if we compare F.M. broadcasting with A.M. broadcasting on long or medium waves, where the large number of transmitters enforces a drastic limitation of the bandwidth and accordingly of the audio spectrum [1]). This limitation however, would disappear if the carrier wavelength in an A.M. transmitter were chosen in the metre waveband, as has been done for the sound channel for TV in England, France and Belgium. One might then expect there to be no difference in sound quality between A.M. and F.M. reception: nevertheless, F.M. is to be preferred for high quality sound broadcasting for the following reasons:

1) With F.M. the signal is contained in the variations in frequency of the carrier wave. All interference which causes variations in amplitude can, in principle, be eliminated and consequently only interference which affects frequency variations can give rise to audible interference after detection. This results in a signal-to-noise ratio which is much better than with A.M., provided that the frequency variation is large enough.
2) The second advantage of F.M. concerns the transmitter. With F.M. the power transmitted is constant, independent of the modulation whilst an A.M. transmitter must have a reserve such that up to 4 times the normal power can be supplied during periods of strong modulation.

As Armstrong showed in 1936, in order to achieve the favourable signal-to-noise ratio potentially available with F.M., the maximum frequency variation ("frequency sweep") must be much greater than the highest audio frequency ²) to be transmitted, viz. 15 000 c/s. By international agreement, the highest permissible frequency sweep for broadcasting is 75 kc/s. With this sweep, a bandwidth of 160 kc/s at the very least is required, so that for F.M. we have to resort to V.H.F. Of the frequency ranges made available for F.M. broadcating by international agreement, that used in Europe is from 87.5 to 100 Mc/s. To profit fully from the wide audio spectrum, clearly the audio frequency section of the set, including the loudspeaker, must give good reproduction of frequencies up to 15 000 c/s. (Of course this is desirable not only for F.M. reception but also for "local reception" of the A.M. transmissions mentioned in note ¹) and for the reproduction of gramophone records.) The problems concerning the audio frequency section of the set will not be discussed here; this article deals with the following matters:

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[1]) It should be noted that some A.M. transmitters do transmit quite high notes (up to 10 000 or even 12 000 c/s), for the sake of those who can receive these transmissions without interference. With this point in mind there are receivers on the market whose bandwidth can be enlarged accordingly.

I) The requirements which a modern F.M. receiver should meet and the way in which these can be realized; also questions which arise in the design of a combined A.M./F.M. receiver;

II) The built-in aerial for F.M. reception.

I. CIRCUITRY AND STRUCTURAL DESIGN

Requirements for an F.M. receiver

In addition to the low frequency section the following points have to be considered in a modern F.M. receiver, viz: the sensitivity, the signal-to-noise ratio, the signal-to-interference ratio, the selectivity and the radiation. These points will now be considered in turn.

Sensitivity

The input signal of the low frequency amplifier, even when receiving from fairly distant transmitters, must be so strong that the volume control never needs to be in its maximum position. If this were necessary, the sound would generally be much too loud and distorted when switched over to A.M. transmitters whose field strength is generally far greater.

Signal-to-noise ratio

The quality of the reception is determined largely by the signal-to-noise ratio at the output of the set. To make this as large as possible, one should aim at:

1) As large a ratio of H.F. signal to noise as possible at the input of the set.
2) As large an H.F. amplification as possible so that the noise arising from the mixer tube and the I.F. tubes is relatively small.
3) An I.F. amplification which ensures that the signal at the control grid of the last I.F. tube is large enough to permit it to be limited (in order that the noise present as amplitude modulation may be largely eliminated).
4) Further limiting of the signal in the detector.

Item 3) above implies the use of 3-stage I.F. amplification. Combined with the large H.F. amplification this makes the total amplification so large that normal criteria for the sensitivity are no longer valid. With A.M. receivers, the usual measure of the sensitivity is the e.m.f. of the aerial which gives an L.F. output power of 50 mW at a certain depth of modulation. With an F.M. receiver, a sensitivity of, for example, 0.4 μV might be found, but at such a low e.m.f. the noise would predominate. Three cases are therefore distinguished as far as the signal-to-noise ratio is concerned according to the strength of the aerial signal. In each one, the aerial e.m.f. required to produce the signal-to-noise ratio given below should not exceed a certain value, according to present standards:

Signal-to-noise ratio dB	Aerial e.m.f. μV	Reception quality
8-10	2	This low quality is considered satisfactory for reception of very weak signals.
26	4	Generally accepted as "good" quality.
40-60	20	Very high quality.

Signal-to-interference ratio

F.M. reception can suffer interference from:

1) Transmitters working on the same central frequency as the transmitter it is desired to receive.
2) Transmitters whose central frequency differs but little from that of the one it is desired to receive.
3) Reception along paths of varying length, such as is often the case in mountainous regions.
4) Sparks from commutator motors and internal combustion engines (without any measures for

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²) See Philips tech. Rev. **8**, 42-50 and 89-96, 1946.

interference suppression, an engine may give rise to a field strength of, say, 500 μV/m at a distance of 10 m).

Limiting again provides an effective means of counteracting this last type of interference in F.M. In general, we require a signal-to-interference ratio of at least 30 dB (for modulation with a frequency sweep of 15 kc/s). As in the case of noise suppression, this implies three I.F. stages.

Selectivity

The carrier frequencies of the European F.M. transmitters are usually 300 kc/s apart. For this separation the selectivity must be at least 200 (i.e. a signal 300 kc/s off-tune must be amplified 200 times less than the in-tune signal). In practice, this selectivity can be achieved only by using three or more stages of I.F. amplification.

Radiation

To avoid interference with nearby F.M. and TV receivers, the oscillator of an F.M. receiver should radiate only very little both at the fundamental frequency and at the first harmonic (which comes within a frequency range used for TV in some countries). Very strict requirements are in force in Germany as regards radiation in the frequency ranges (87.5-100) + 10.7 ≈ 98-111 Mc/s (10.7 Mc/s is the intermediate frequency) and 196-222 Mc/s.

F.M. receiver circuits

We shall now examine how the requirements outlined above are met in Philips F.M. receivers.

The H.F. amplifier

The input of the H.F. section is matched to an aerial impedance of 280 ohms (see part II of this article); the characteristic impedance of the feeder between aerial and set also has this value. Theoretically, the noise factor should be 2 but for various reasons it is approximately 3.5 in practice, a value which is easy to maintain in mass-production.

In order to keep the noise from the tubes as small as possible, triodes are used in the H.F. section because, unlike pentodes, they are free of distribution noise. A tube that is frequently used is the double triode ECC 85, one half of which functions as an H.F. amplifier and the other half as an oscillator-mixer tube.

A disadavantage of triodes, however, is that the grid-anode capacitance C_{ag} is fairly large. When using a grounded-cathode circuit, it would be necessary to neutralize in order to prevent the output circuit reacting unduly on the input circuit via C_{ag}, but this is not so easy on account of the rather large frequency range (87.5-100 Mc/s). Neutralizing presents no problem in the grounded-grid circuit, for this involves the anode-cathode capacitance C_{ak}, which is much smaller than C_{ag} and moreover has less effect because the input impedance of the tube in this circuit is very low, viz. smaller than $1/S$, where S is the mutual conductance. In the ECC 85, with $S = 5$ to 6 mA/V, the input impedance in a grounded-grid circuit is roughly 150 ohms. This circuit however has the disadvantage of low amplification (ratio of anode H.F. voltage to aerial e.m.f.).

A favourable compromise is to be found in the "Zwischenbasis" circuit (*fig. 1a*) which is intermediate between the grounded-cathode and grounded-grid circuits. Neither the cathode nor the grid are common to both the input and circuits; instead, an intermediate point, for example a tapping on the coil S_2, is made common. As the tapping is moved downwards (in this figure) the circuit tends towards the grounded-cathode circuit; moved upwards, we approach the grounded-grid circuit. By a suitable choice of the position of the tapping sufficient amplification can be obtained from only one tube and the input impedance of the tube can be made considerably higher than $1/S$ (and the damping in the input circuit small in consequence); also the neutralizing is not very critical. The condition for neutralizing with a given tapping point is that there should be a specific ratio between the grid-anode capacitance and the total anode-cathode capacitance (see fig. 1*b*). In our

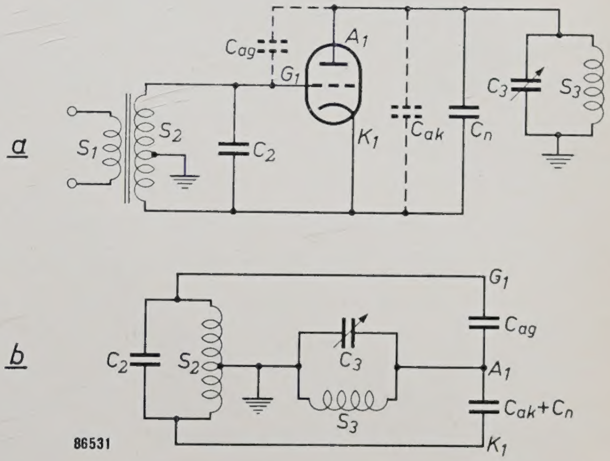

Fig. 1. *a*) Diagram of the H.F. amplifier in "Zwischenbasis" circuit. *b*) Equivalent circuit. S_1 primary coil. S_2-C_2 secondary circuit, tuned to a fixed frequency at the middle of the frequency range (94 Mc/s). A_1 anode, G_1 grid and K_1 cathode of one half of the double triode ECC 85. S_3-C_3 output circuit. C_{ag} grid-anode capacitance. C_{ak} anode-cathode capacitance. C_n neutralizing capacitance. The power supplies are not shown.

case the correct value of this ratio is obtained by adding a certain neutralizing capacitance C_n to the stray capacitance C_{ak} already present. Once the right values had been found, it was possible to use a fixed tapping and a fixed neutralizing capacitor in mass-production, with no subsequent adjustment.

The mixer stage

The H.F. amplification stage is followed by the frequency-changer stage consisting of oscillator and mixer. As already stated, 10.7 Mc/s is chosen as the intermediate frequency. As regards the mixer tube the most important properties are minimum noise and large conversion coductance. The triode is the best valve to meet these requirements and as mentioned earlier, an ECC 85 is often used. Owing to the large H.F. amplification of the previous stage the amount of noise from the mixer tube is negligible. Since a triode has only one grid, a so-called additive mixing circuit has to be used, i.e. a circuit in which the two signals to be mixed are applied to the same grid.

Fig. 2 shows the layout of the H.F. stage and the mixer stage. Two features of the circuit will now be discussed: the first concerns the damping of the first I.F. band pass filter and the second concerns the stray radiation.

The plate resistance of the mixer tube measured on D.C. amounts only to about 20 000 ohms, which implies a fairly strong damping in the band-pass filter and, as a result, low amplification. In addition, an I.F. voltage passes to the grid via the grid-anode capacitance, causing the effective plate resistance to become even lower, viz. about 5000 ohms. However, the bridge circuit shown in fig. 2*b* causes another I.F. voltage to appear at the grid such that the effective plate resistance is increased to a value at which the damping is small enough. The bridge circuit comprises four capacitances (regarding coils S_3, S_4 and S_5 as short-circuits for I.F. currents). Of these four capacitances C_{ag} is given by the tube and C_6 and C_7 are for other reasons also fixed within certain limits. Typical values in practice are: $C_{ag} = 2$ pF, $C_6 = 20$ pF and $C_7 = 200$ pF. With these values the bridge would be balanced if $C_5 + C_3 = 2000$ pF. By choosing another value for $(C_5 + C_3)$, the I.F. voltage appearing across the

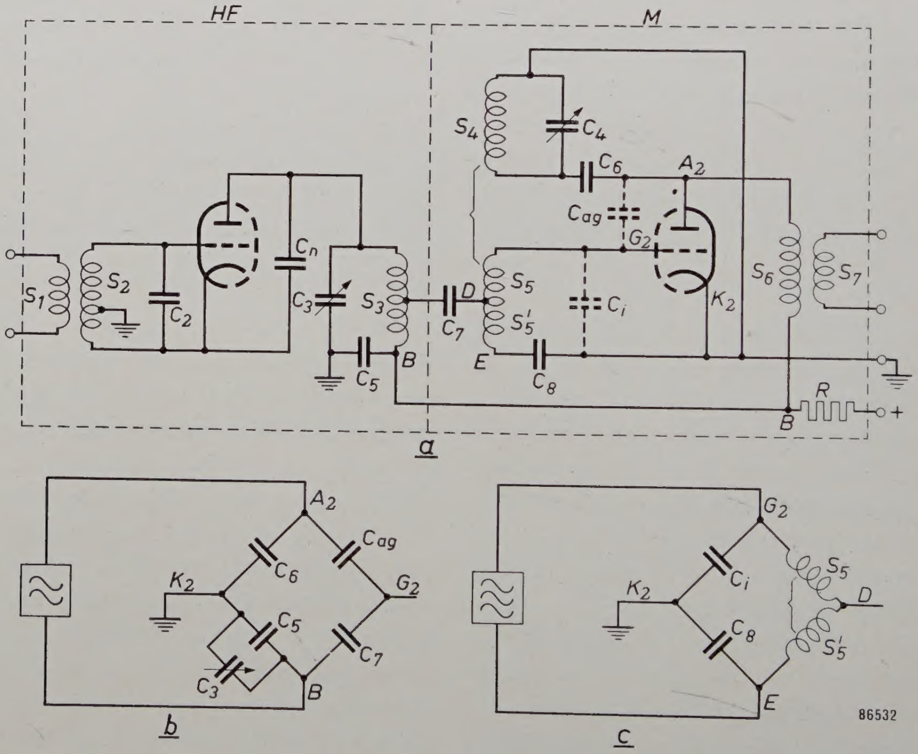

Fig. 2. *a*) Simplified diagram of high frequency amplifier *HF* and mixing stage *M*. For S_1, S_2, S_3, C_2, C_3 and C_n see figure 1. A_2, G_2, K_2 are electrodes of the other half of the double triode ECC 85. S_4-C_4 oscillator circuit. $S_5 + S'_5$ feed-back coil. S_6-C_6 primary circuit of the first I.F. band-pass filter. S_7 coupling coil. C_i input capacitance.
b) Bridge circuit in circuit (*a*) which, at the correct value of C_5, causes a reduction in the damping of the band-pass filter.
c) Second bridge circuit, incorporated in (*a*) which, at the correct value of C_8, ensures that no voltage at the oscillator frequency appears at point *D*, so that there is no radiation.

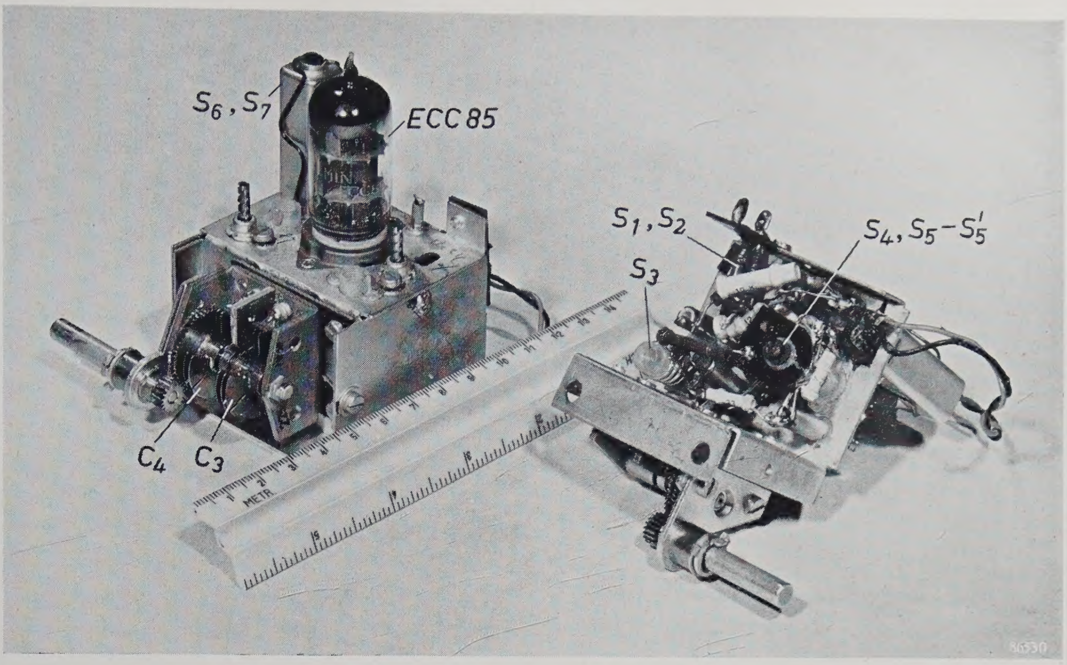

Fig. 3. One form of the circuit of figure 2*a*, seen from above and below. The letters refer to the corresponding components in fig. 2.

diagonal K_2-G_2 can be varied; with $C_5 \approx 900$ pF, this voltage has the correct magnitude and sign. The capacitance C_3 of the tuning capacitor is adjustable from 2.5 to 12.5 pF, and this range is so small compared to C_5 that the required condition is maintained over the entire tuning range.

A second bridge circuit is incorporated in the circuit of fig. 2*a*, to counteract radiation. This is shown in fig. 2*c* and is formed by the two sections of the reactance coil $S_5 + S_5'$, the input capacitance C_i of the oscillator tube and the capacitor C_8. The bridge is balanced with C_8 so that the diagonal K_2-D remains free of oscillator voltage and this voltage can therefore not reach the aerial via D and the H.F. amplifier. The self-inductances, mutual inductances and capacitances in the bridge circuit are such that the balance of the bridge is theoretically independent of the frequency. Stray self-inductances somewhat impair this independence of frequency in practice, but in the range within which the oscillator frequency varies the balance holds satisfactorily.

In this way and with very short wiring, the circuit in figure 2*a* can produce a total amplification of 200× (ratio of I.F. voltage on control grid of first I.F. tube to aerial e.m.f.). *Fig. 3* shows two views of such a H.F. section.

The I.F. amplifier

As already mentioned, there are a number of reasons for using not less than three I.F. stages in an F.M. receiver, e.g. to achieve sufficient selectivity and especially sufficient amplification in order that undesirable amplitude variations (noise and interference) can be eliminated by limiting.

In the first I.F. stage, use can be made of the heptode section of the triode-heptode (ECH 81) which does service as a mixer tube in A.M. reception. The triode section remains out of use for F.M. reception. In the second stage, the best tube to use is the EF 89 pentode since with this type, the ratio S/C_{ag}, which determines the amplification, is largest. The last stage must function as a limiter particularly for strong signals (with weak signals it is mainly the detector which functions as a limiter). The EF 85 tube has the most favourable characteristics for this. This tube is used in such a way that limiting is achieved by grid current; this prevents unwanted amplitude variations in the anode current.

Automatic volume control is arranged by taking a control voltage from between the second and third I.F. tubes and feeding this back to the control grid of the first I.F. tube. In this way, the automatic volume control has no direct influence on the limiting effect of the third I.F. tube.

The detector

The most modern F.M. detector is the "ratio detector", which has the property of not responding to amplitude variations of the signal and therefore works as a limiter [3]). The advantages of this detector over the Foster and Seeley detector originally developed in America[4]) are as follows; 1) better limiting for weak signals, and 2) less interstation noise, i.e. the noise the receiver produces when it is not well-tuned (for example while it is being retuned to another station).

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[3]) With the addition of one or two circuit elements, this detector can easily be converted into an A.M. detector; this is particularly convenient in the case of dual or multi-standard TV receivers. See Philips tech. Rev. **17**, 168, 1955/1956 (No. 6), fig. 12.

The basic circuit of the ratio detector is reproduced in *fig. 4a*.

The following discussion is intended to show how this circuit works as an F.M. detector (and not as an A.M. detector) without, however, claiming to be a complete explanation.

Consider first the property of band-pass filters that the difference in phase between the primary and the secondary voltage increases with frequency and at the resonance frequency is equal to 90°.

In fig. 4*a* L_1-C_1-L_2-C_2 form a band-pass filter whose resonance frequency is equal to the central intermediate frequency f_c. When there is no frequency modulation (frequency f equal to f_c), fig. 4*b* applies: the secondary voltages $\mathbf{V}_1$ and $\mathbf{V}_2$ are phase-shifted + 90° and — 90° respectively with respect to the primary voltage $\mathbf{V}_0$ (letters in the bold type represent vectors).

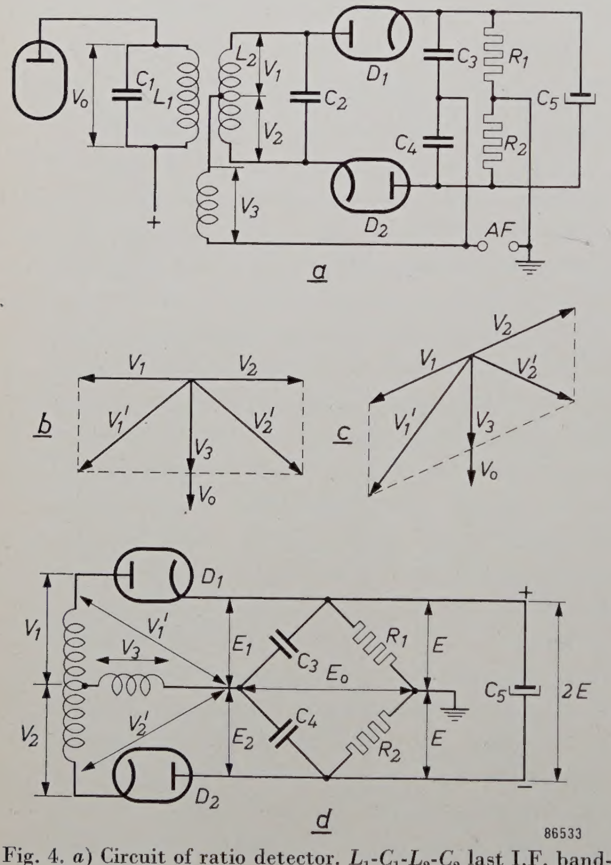

Fig. 4. *a*) Circuit of ratio detector. L_1-C_1-L_2-C_2 last I.F. band-pass filter. *AF* terminals across which the audio frequency voltage appears.
b) Vector representation of the I.F. voltages $\mathbf{V}_0$, $\mathbf{V}_1$, $\mathbf{V}_2$ and $\mathbf{V}_3$ (see (*a*)) in the absence of any frequency modulation. The sum voltages $\mathbf{V}_1'$ and $\mathbf{V}_2'$ which are rectified are of equal magnitude.
c) Same as (*b*), but with a certain frequency deviation. $\mathbf{V}_1'$ and $\mathbf{V}_2'$ are no longer equal.
d) Same as (*a*) but showing the direct voltages E_1 and E_2 (varying at audio frequency), the audio frequency output voltage E_0 and the direct voltage $2E$ (assumed constant) across the electrolytic capacitor C_5.

To $\mathbf{V}_1$ and $\mathbf{V}_2$ a voltage $\mathbf{V}_3$ is added which is always in phase with $\mathbf{V}_0$. The sum voltages $\mathbf{V}_1'$ and $\mathbf{V}_2'$ are each rectified separately (diodes D_1 and D_2) so that the direct voltages E_1 and E_2 appear across the smoothing capacitors C_3 and C_4 respectively. We then have:

$$\frac{E_1}{E_2} = \frac{V_1'}{V_2'}\,, \quad \dots\dots\dots\dots \quad (1)$$

where V_1' and V_2' are the moduli of $\mathbf{V}_1'$ and $\mathbf{V}_2'$. Since $\mathbf{V}_1$ and $\mathbf{V}_2$ are equal in magnitude ($V_1 = V_2$), when there is no frequency modulation $V_1' = V_2'$, $E_1 = E_2$ and the output voltage E_0 is nil (fig. 4*d*).

If there is frequency modulation, the phase angle between $\mathbf{V}_1$ and $\mathbf{V}_0$ is alternately larger and smaller than 90°. Fig. 4*c* gives the state of affairs for a particular frequency deviation: V_1' and V_2' are now no longer equal, neither are E_1 and E_2 and there is consequently a certain output voltage E_0.

Since the voltage across the electrolytic capacitor C_5 can be regarded as constant (value, say, $2E$), for short periods, a constant voltage E appears across each of the two equal resistors R_1 and R_2. We then have:

$$\left.\begin{aligned} E_1 &= E - E_0 \\ E_2 &= E + E_0 \end{aligned}\right\} \quad \dots\dots\dots\dots \quad (2)$$

From (1) and (2), we may write:

$$E_0 = \frac{V_2' - V_1'}{V_2' + V_1'}\,E\,. \quad \dots\dots\dots\dots \quad (3)$$

With a correct choice of component values, E_0 varies almost proportionally to the frequency deviation.

If the input signal V_0 of the detector changes in amplitude, for example by a factor a, then V_1' and V_2' also change by this factor, so that the ratio $(V_2' - V_1')/(V_2' + V_1')$ remains constant (hence the name "ratio detector"). If the change in V_0 is rapid in comparison with the large time-constant $(R_1 + R_2)C_5$, then E remains practically constant and hence from (3) E_0 also remains constant. In other words, any amplitude modulation which might arise in V_0 is not manifested in E_0. Because the diodes are not ideal, the suppression of amplitude modulation is not complete.

In F.M. transmissions "pre-emphasis" of the high notes is used, i.e. the frequency sweep of the transmitter for an audio signal of constant amplitude is made to increase with frequency. At the receiver end, an equal "de-emphasis" must be applied to restore the original balance. The advantage of this method is to be found in the fact that with the de-emphasis the noise is attenuated relatively more than the signal and this results in reproduction with an improved signal-to-noise ratio. The pre-emphasis is achieved by using a high-pass *LR*-filter before the modulation stage; for de-emphasis in the receiver, a low pass *RC*-filter is incorporated after the detector stage. Pre-emphasis and de-emphasis compensate each other if both filters have the same time constant. On the continent of Europe this constant is fixed at 50 μsec by agreement.

Structural design

Circuit requirements as outlined above impose certain restrictions on the structure of an F.M.

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[4]) See Philips tech. Rev. **8**, 48, 1946, fig. 10.

receiver in particular with regard to the location of certain components and the control knobs.

It is not advisable to mount the A.M. and F.M. tuning condensers on a single spindle; the component locations would then preclude the use of the shortest wiring in the F.M. section. The simplest solution is for each condenser to be driven by its own knob. If the set is provided with push-buttons (on/off switch, waveband selection, etc.) the set can be left tuned in to both an A.M. transmitter and an F.M. transmitter; the set can then be switched on at either of these stations by pressing the appropriate button without needing to tune again. To save space on the front of the set, the two tuning knobs can be arranged concentrically (*fig.* 5). The condensers can be turned via the usual cord-drive.

It is also possible to use only one tuning knob which can be coupled with the A.M. or the F.M. tuning unit as desired. The mechanical coupling of the knob is done automatically when the required wave band is selected by means of buttons (*fig.* 6). This type of tuning control imposes considerable demands on the reproducibility of the adjustment since only a slight mechanical inaccuracy can adversely affect the sound quality and the freedom from interference.

Pre-tuning, with push-button selection might also be applied to more than two transmitters, but of course for each extra transmitter a separate manual tuning arrangement is necessary, and a corresponding push-button. In the more de luxe models one might use motor-driven tuning, the motor being stopped automatically at the desired point by pressing an appropriate button. Features such as this, however, are beyond the scope of this article.

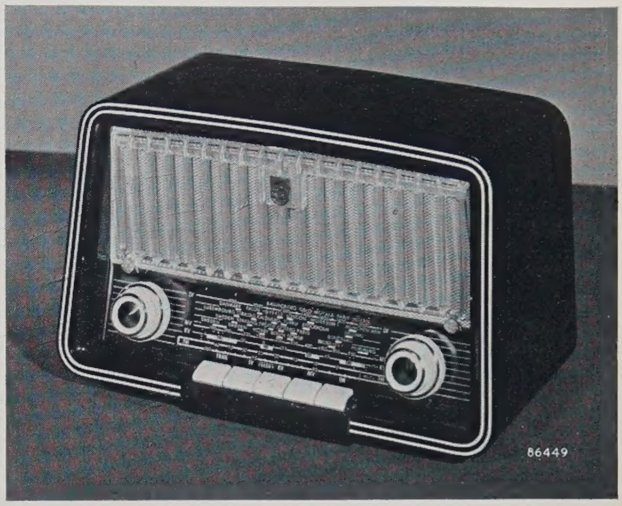

Fig. 5. Receiver with separate tuning knobs for AM and FM (type BX 253 U). In the right is a double knob for tuning; the front knob is used for AM and the rear for FM. The left-hand knob is also double; the front knob is used for the volume and the rear one for the tone control. The buttons in the centre are (from left to right); on/off switch, long wave, short wave, medium wave, and F.M. reception. For gramophone reproduction, the long and the short wave buttons are pressed in together.

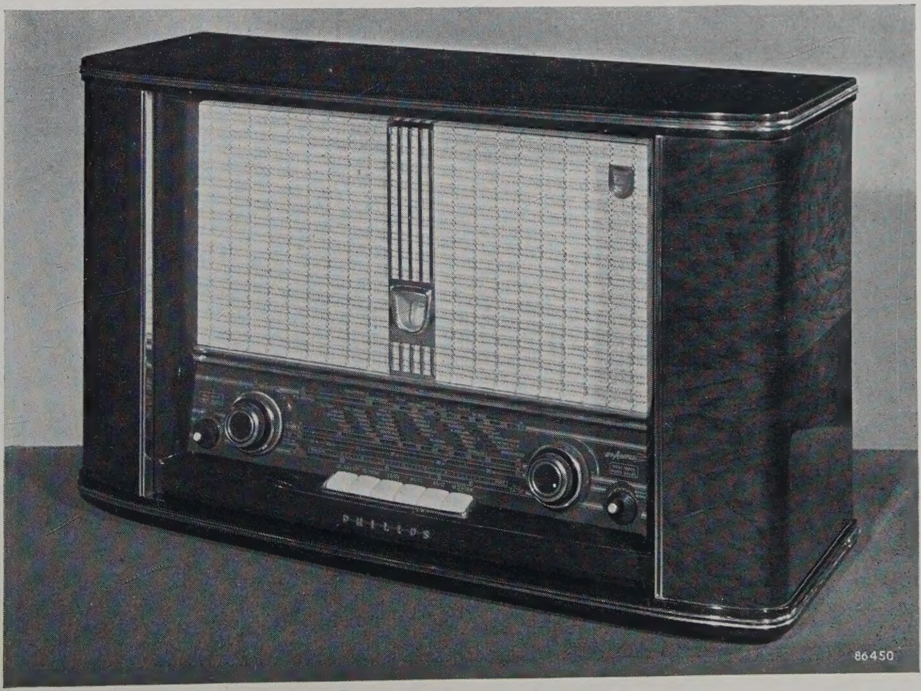

Fig. 6. Receiver (type BX 653 A) with a common tuning knob (the large knob on the right which appears to be double but is, in fact, single). This controls the A.M. tuning when an A.M. button is pressed in and the F.M. tuning when the F.M. button is pressed in. The knob on the left is double; the front knob is the volume control and the rear one adjusts the orientation of the "Ferroceptor" aerials. The knobs on the extreme left and right are independent tone controls for low and high notes.

II. THE BUILT-IN F.M. AERIAL

Many listeners are unable for various reasons to instal a good outside aerial. For short-wave reception (10-100 m, A.M.) this is not necessary since the small plate aerial with which the majority of sets are now equipped ensures good and practically interference-free reception on these wavelengths nearly everywhere. In the case of medium and long waves, however, the situation is different. In the absence of an outside aerial, use is often made of an indoor aerial. In many cases this gives sufficiently strong reception but especially in towns, interference may be troublesome. Improvements in long and medium wave reception have been effected by equipping sets with inductive aerials for these wavebands (either frame aerials with one or two turns or "Ferroceptor" aerials [5]).

For receivers incorporating a frequency modulation unit it is convenient to equip them with a built-in aerial for F.M. reception as well, especially as with ultra short waves the circuit parameters and matching require some care to get the best results.

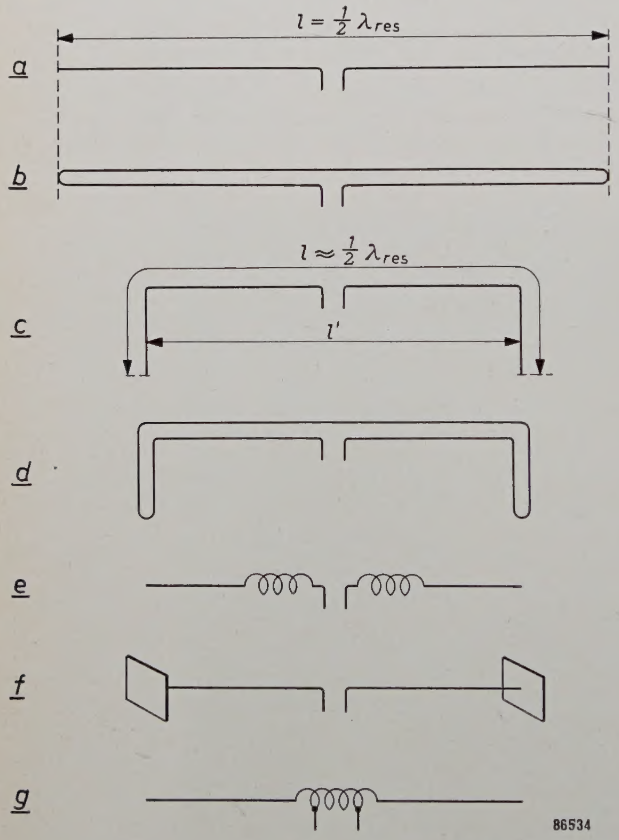
$l = \frac{1}{2}\lambda_{res}$
a
b
$l \approx \frac{1}{2}\lambda_{res}$
c
l'
d
e
f
g
86534

Fig. 7. Forms of dipole aerials. *a*) Normal dipole; length l equal to half the resonance wave length λ_{res}. *b*) Folded dipole. In order to fit the dipole into a cabinet of length $l' < l$, it can either be bent (*c*, *d*), or provided with two coils (*e*), two end plates (*f*) or one coil (*g*).

This F.M. aerial — a dipole — can also be used for reception on amplitude modulated short waves. Many modern radio sets therefore incorporate three built-in aerials, viz. two inductive aerials and a dipole.

A dipole in its simplest form is shown in *fiig. 7a* Electrically speaking it may be replaced by an equivalent generator with a certain internal impedance. The magnitude and the nature of this impedance are important for the matching, which will be dealt with presently. In the case of fig. 7*a*, the impedance has the same character as a resonant circuit of self-inductance, capacitance and resistance in series. At resonance (wavelength $\lambda_{\text{res}} = 2 \times$ length l) the impedance is a resistance of roughly 70 ohms; at higher frequencies it is inductive in character and at lower frequencies capacitative. The folded dipole (fig. 7*b*) behaves similarly, only here (the leads having the same diameter) the resonance resistance is 4 × as large, i.e. roughly 280 ohms.

Since the wavelength in the F.M. band is three metres or more and a dipole such as in fig. 7*a* or *b* must accordingly be about 1.5 metres long, it cannot be housed in the cabinet of a radio set without its shape being modified as shown in figures 7*c* and *d*. It appears that this bending of the aerial hardly alters the resonance frequency but that the radiation resistance (which forms the greater part of the resonance resistance) decreases considerably. In the case of a cabinet whose largest dimension is between 20″ and 30″, the resonance resistance is 20 to 30 ohms with the normal dipole (fig. 7*c*) and 100 to 120 ohms with the folded dipole (fig. 7*d*).

It is possible to make do with a dipole shorter than $\frac{1}{2}\lambda_{\text{res}}$ by adding self-inductance or capacitance. In fig. 7*e* the self-inductance is increased by two coils and in fig. 7*f* the capacitance is increased by two end plates. These kinds of aerial have almost the same characteristics as a bent dipole (fig. 7*c*). If in fig. 7*e* the two coils are replaced by one coil having a self-inductance twice as large (fig. 7g), the required impedance may be achieved by means of tappings on the coil.

In Philips sets a bent folded dipole is used (fig. 7*d*). The length is so chosen that the resonance occurs in the middle of the frequency range, i.e. at 94 Mc/s.

Matching of aerial and receiver

The signal at the input terminals *c-d* of the receiver (*fig. 8*) is strongest if the aerial impedance $Z'_{\text{a}} = R_{\text{a}} + \text{j}X_{\text{a}}$ is matched to the impedance $Z_{\text{i}}' = R_{\text{i}}' + \text{j}X\text{i}'$ which the aerial "sees" at the input *a-b* of the lead connecting it to the set, i.e. if

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[5] H. Blok and J. J. Rietveld, Philips tech. Rev. **16**, 181-194, 1954/1955 (No. 7).

$R_a = R_i'$ and $X_a = -X_i'$. The impedance Z_i' generally differs from the input impedance $Z_i = R_i + jX_i$ between the terminals *c-d* of the receiver itself in that the feeder cable functions as an impedance transformer. Only if Z_i is equal to the characteristic impedance ζ of the feeder over the whole frequency range, is $Z_i' = Z_i$.

A folded dipole used as an outside aerial (with R_a = 280 ohms) can be properly matched by connecting it via a suitable lead (ζ = 280 ohms) to a receiver with an input impedance at resonance likewise of 280 ohms.

With a bent folded dipole inside the set (fig. 7*d*), we have $R_a \approx$ 100-120 ohms whilst R_i = 280 ohms. The short connecting line has to make as good matching as possible. In order to check the matching we can profitably use a measuring technique [6]) by means of we obtain values of the components of Z_a and Z_i'. It is then possible to see whether the conditions for matching: $R_a = R_i'$ and $X_a = -X_i'$, are satisfactorily met. If this is not the case, the measurement may be repeated with a connecting line of other dimensions: by such a process of trial and error, satisfactory matching can be achieved.

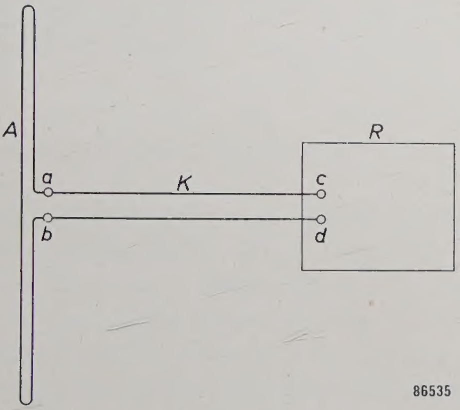


Fig. 8. Folded dipole *A* connected by a feeder *K* to the receiver *R*.

On the last page of the article quoted in [6]) a number of variants of the measuring technique described are given. The last of these has been used in the present work: a short description of the measuring arrangement will now be given.

In *fig. 9*, *G* is a signal generator with a known frequency *f* which is adjustable from 87.5 to 100 Mc/s. The output voltage V_{gen} is continuously adjustable and its relative value is known. The generator is connected in series with resistances R_0 to a transmission line roughly 60 m long. The value R_0 is so chosen that $2R_0$ plus the internal resistance of the generator is equal to the characteristic impedance of the transmission line. The other end of the latter is either short circuited or connected to the impedance *Z* which is to be measured (i.e. the impedance Z_a of the dipole or the input impedance Z_i' of the feeder *K* in fig. 8). At the generator end of the transmission line there is a detector *D* and an electronic millivoltmeter *mV*. Since V_{gen} is modulated up to a particular depth in amplitude, the detector voltage V_{det} read off from the meter *mV* is a measure for the H.F. voltage at the beginning of the transmission line.

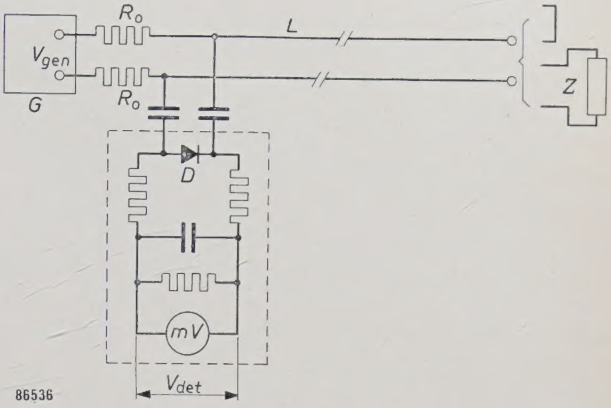
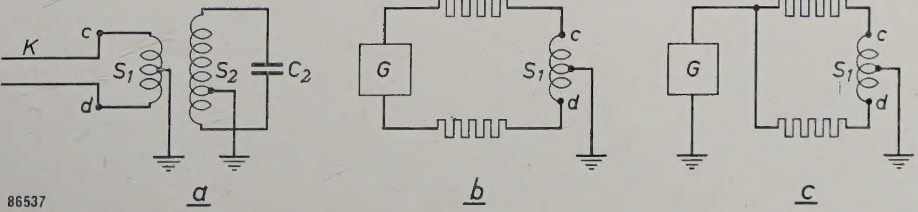


Fig. 9. Circuit for measuring impedances. *G* is a signal generator with adjustable frequency and voltage, amplitude-modulated. *L* is a long transmission line. *Z* is the impedance to be measured, *D* the detector and *mV* the electronic millivoltmeter.

Measuring the sensitivity

For measuring the sensitivity of a built-in dipole a millivoltmeter is connected in the limiting circuit or in the detector circuit of the receiver. This meter is calibrated in H.F. voltage at the input of the receiver.

The receiver is set up as far away as possible from objects which might reflect the waves. A transmitting dipole is placed not less than 30 metres away so that its plane of symmetry coincides with that of



a *b* *c*

Fig. 10. *a*) Input of an F.M. receiver showing the feeder *K* of the built-in dipole. S_1, S_2 and C_2 as in fig. 1. *b*) and *c*). Checking for symmetry. *G* signal generator. The millivoltmeter connected in the set should show a deflection no greater in case (*c*) then it was in (*b*), when the generator voltage is made 10 times smaller than in the latter case.

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[6]) J. C. van den Hoogenband and J. Stolk, Philips tech. Rev. **16**, 309-320, 1954/1955 (No. 11).

the receiving aerial. The transmitting dipole produces a field whose strength at the receiver is known. The magnitude of the input signal resulting from this field strength is then read from the meter in the receiver.

The possibility of asymmetry at the input of the receiver has to be considered. The two wires of the feeder K (*fig. 10a*; the dipole is assumed to be not yet connected) function themselves rather like an aerial and supply in-phase signals to points c and d. When the input it absolutely symmetrical these signals cancel each other out, but when asymmetrical, as they are in general, a certain residual signal remains, say V_1. Let the desired signal provided by the dipole be V_2. According to the way in which the dipole is connected (connection is done by means of interchangeable plugs), the total signal becomes the sum of the vectors of V_1 and V_2 or V_1 and $-V_2$. The difference between the two sums and therefore the difference in sensitivity can be considerable.

For this reason it is desirable to check the symmetry beforehand. To do this a signal generator is connected in turn as shown in figs. 10*b* and *c*. In the latter case the meter would show zero deflection if there was perfect symmetry. The symmetry attained may be regarded as satisfactory if the deflection obtained in case *c* is less than that which would be obtained in case *b* with a 10× smaller input signal.

Summary. The principal electrical properties required of an F.M. receiver concern the sensitivity, the signal-to-noise ratio, the signal-interference ratio, the selectivity, the radiation and the quality of the audio-frequency section. These requirements have led to the design of sets incorporating the following features: single stage H.F. amplification with one half of a double triode ECC 85 in "Zwischenbasis circuit", a frequency changer stage using the other half of the ECC 85, three stage I.F. amplification (with the heptode section of an ECH 81, a pentode EF 89 and a pentode EF 85), a ratio detector and an audio frequency section (not described in this article) with good response up to 15 000 c/s. Two bridge circuits are incorporated in the mixing stage, the one ensures that the signal attenuation in the first I.F. band pass filter is small and the other prevents radiation. Sets designed for both A.M. and F.M. reception are equipped either with two independent tuning knobs or with one tuning knob which is coupled to the appropriate tuning unit by means of press-buttons.

The second part of this article deals with the built-in F.M. aerial: a folded dipole, the ends of which are bent to enable it to be housed in a radio cabinet. The bent ends change the aerial impedance from 280 ohms to 100-120 ohms. A short feeder of suitable length ensures matching with the input impedance of the set, which, with an eye on the use of an external dipole, is kept at 280 ohms. A method for checking the matching is briefly described.

A RECEIVER FOR THE RADIO WAVES FROM INTERSTELLAR HYDROGEN

II. DESIGN OF THE RECEIVER

by C. A. MULLER *).

522.6:621.396.9

Under the auspices of the Netherlands Foundation for Radio Astronomy, extensive observations are being made of the 21 cm radiation emitted by interstellar hydrogen. This radiation, which has the character of noise, is extremely weak and special measures are required to be able to distinguish if from the very much stronger noise inherent in the receiver. The first part of this article outlined the principles on which these investigations are based. The second part, printed below, describes the receiver circuits employed, with special reference to the function and design of certain components.

Principle of the receiver

The receiver about to be described was designed for investigations on the 21 cm wavelength (1,420 Mc/s) region of the radio spectrum from the Galatic System. As discussed in part I of this article [1]), there is a spectral line at this wavelength, which is radiated by clouds of hydrogen atoms in interstellar space.

The radiation from interstellar hydrogen has the character of noise, that is to say its intensity varies randomly with time. Considered as a function of frequency, the average intensity shows a marked maximum around 1,420 Mc/s. The form of the spectral line, i.e. the line profile to be measured, depends upon the bearing from which the radition is received A typical profile is shown schematically in *fig. 1.* The origin of these profiles and their significance were outlined in part I.

The principle of the measurement is as follows. The receiver, which is a superhet with three mixing stages, is switched to and fro at a rate of 400 times per second between two tuned positions between which a constant difference is maintained of $\Delta f =$ 1080 kc/s. These tuned positions are together displaced in such a way that one of them always falls outside the frequency band occupied by the spectral line (the width of the line is generally less than 1 Mc/s) and the difference in output voltage at both frequencies is recorded. In this way the complete profile of the spectral line is traversed. As the signal at the one frequency is always somewhat stronger than at the other, the output voltage after rectification displays a square waveform with a frequency of 400 c/s. Its amplitude is a measure of the difference in the noise power received at both frequencies.

Since the receiver is tuned to a frequency outside the spectral line (the comparison frequency) during each half period of the measurements, the available information over the line profile is received for only half the time. The accuracy of the measurement is thus reduced by a factor $\sqrt{2}$. (We shall return to this point later.)

In the actual receiver this drawback has been overcome by using a parellel arrangement of two I.F. amplifiers, tuned to frequencies which differ from each other by the above-mentioned $\Delta f =$ 1080 kc/s. Two measurements can now be carried out simultaneously; when the receiver is in the one tuned position, the first I.F. channel amplifies the comparison frequency f_2 while the second (owing to the frequency difference) amplifies the spectral line

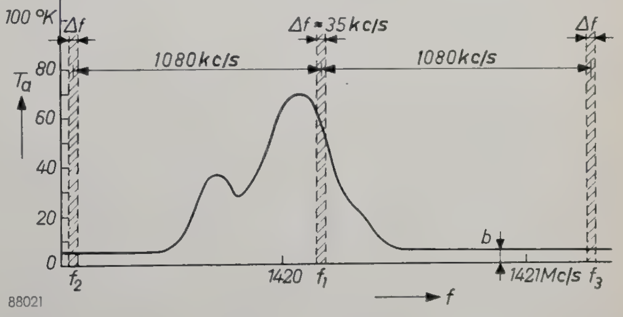

Fig. 1. Schematic diagram of a typical profile of the 21 cm line radiated by interstellar hydrogen. The noise intensity received is expressed in "antenna temperature", T_a. Continuous background radiation is present (*b*), the temperature of which is normally less than 10 °K. The receiver is switched to and fro between the frequencies f_1 (on the spectral line) and f_2 and f_3 (adjacent to it, on opposite sides) The bandwidth in each case is about 35 kc/s. All three frequencies are simultaneously displaced through the spectrum, a constant frequency interval of 1080 kc/s being maintained between them.

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*) Director of the observations group of the Netherlands Foundation for Radio Astronomy.

[1]) C. A. Muller, A receiver for the radio waves from interstellar hydrogen, Part I, Philips tech. Rev. **17**, 305-315, 1955/56.

frequency f_1. In the other tuned position, however, the first I.F. channel amplifies frequency f_1, and channel 2 amplifies a second comparison frequency f_3, which, like f_2, differs from f_1 by $\Delta f = 1080$ kc/s, but in the opposite direction (see fig. 1). All three frequencies are together slowly displaced, so that f_1 again traverses the spectral line, while both f_2 and f_3 remain always outside it. After rectification, the output voltages of both channels each display a square waveform, and these two squarewave voltages, which are in anti-phase, are then added together in a push-pull amplifier and their combined amplitude measured. In this way, therefore, we are able to observe the spectral line uninteruptedly and to compare the received power with the average power at the two frequencies, f_2 and f_3.

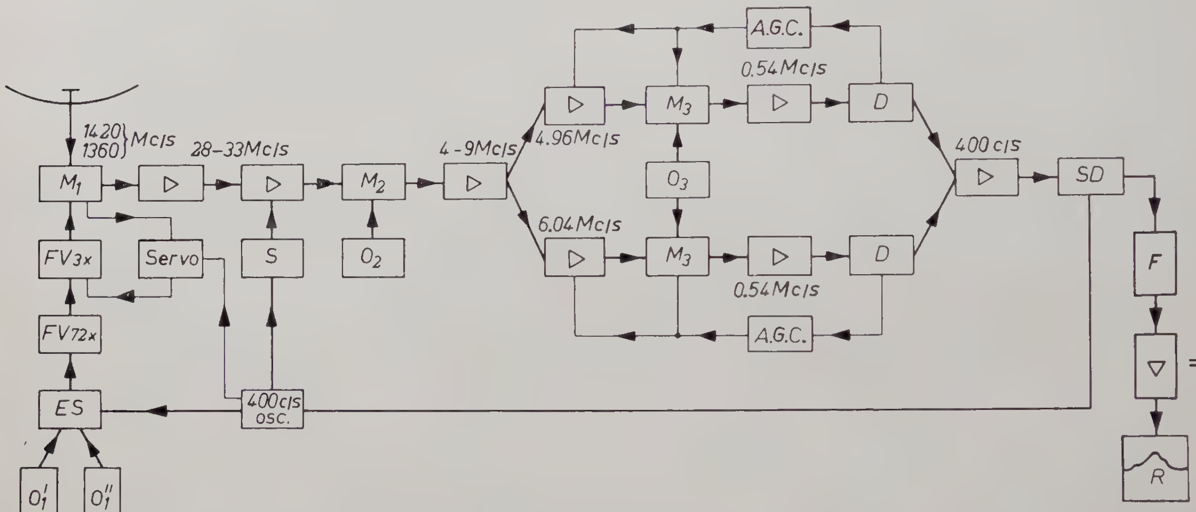

Fig. 2. Block diagram of the receiver. The units joined by thick lines constitute the receiver proper; those joined by thin lines represent control circuits. O'_1 and O''_1 are the crystal oscillators of the first mixer stage M_1; O_2 and O_3 are the oscillators of the second and third mixers, M_2 and M_3 respectively. *FV* frequency multipliers, *ES* electronic switch, *D* detectors, *AGC* automatic gain control, *SD* synchronous detector, *F* low pass filter, *R* strip-chart recorder.

A receiver designed on these principles must satisfy two stringent requirements. In the first place, it must be completely identical in both switching positions, as otherwise a 400 c/s square-wave voltage might appear even in the absence of a signal from the Galaxy. Secondly the switching process must have no effect on the measurement, that is to say, it must not cause the appearance in the output voltage of any component with a frequency of 400 c/s. These requirements principally determine the design of the receiver.

Fig. 2 shows the block diagram of the receiver. The incoming signal, which contains a wide-range of frequencies, is fed to the first mixer stage M_1, where it is combined with an oscillator signal with a frequency of about 1390 Mc/s. The difference frequency is fed to a wide-band amplifier (bandwidth approximately 5 Mc/s) operating in the region of 30 Mc/s. Thus, the amplified signal originates from those components of the incoming signal which contain frequencies in bands of 5 Mc/s around the spectral line frequency of 1420 Mc/s and the image frequency of 1360 Mc/s. The 30 Mc/s signal is combined in a second mixer stage (M_2) with an oscillator signal slowly changing in frequency between 34 and 39 Mc/s. The signal emerging from this stage, which consists of frequencies of a few Mc/s, is amplified and fed to two selective amplifiers in parallel operating at frequencies of 6.04 and 4.96 Mc/s. This produces the two I.F. channels mentioned above differing from each other by 1·080 Mc/s. The output signals of both amplifiers are combined in two third mixer stages (M_3) with the 5·50 Mc/s signal from a common oscillator. The resultant difference-frequency in both channels is therefore 0·54 Mc/s. The two amplifiers used for subsequently amplifying the signals in both channels have the same band-pass response, characterized by the narow bandwidth of 35 kc/s. These are the amplifiers which restrict the signals to two narrow bands within the total frequency spectrum received, that is to say, two bands 1·080 Mc/s apart in the region of 1420 Mc/s, and two similar bands around the image frequency. (The latter are of no further interest and, in any case, the signal received in these bands is extremely weak).

The exact frequencies in the centre of the bands (the tuning frequencies) are determined by the oscillator frequencies of the first and second stages. As the frequency of the second oscillator is slowly varied, the tuning frequencies vary likewise; they move slowly over several Mc/s through the spectrum thereby traversing the spectral line.

The oscillator signal for the first mixer stage is taken alternately from two oscillators, O_1 and O_1'', which differ in frequency by 1·050 Mc/s, and which are alternately gated by a separate 400 c/s oscillator. The result is that both tuned bands of the receiver are switched to and fro at a rate of 400 times per second between two positions lying 1·080 Mc/s apart so that, in turn, one of these frequencies lies within the spectral line while the other is adjacent to it. Since the amplitude of the ultimate square-wave voltage depends upon the gain in both 0·54 Mc/s amplifiers, automatic gain control is applied in both channels and the average output held constant.

Finally, after detection in both channels, the absolute values of the 400 c/s signals obtained are added in a push-pull amplifier, and the summed voltage is fed to a synchronous voltmeter. This measures and records the fundamental of the 400 c/s squarewave voltage which appears at the output of the push-pull amplifier.

Apart from the radio receiver itself, the equipment also contains many semi-automatic calibrating and control circuits.

We shall now consider the various parts of the receiver individually.

H.F. section and crystal mixer

First mixer

The antenna consists of a half-wave dipole with a flat, circular reflector mounted at the focus of a paraboloid of 7½ m diameter (see I). It is connected to the first mixer by a coaxial feeder approximately two metres long; the outer conductor is of rigid copper tubing which also serves to support the dipole. The line is air-insulated apart from spacers of foamed-plastic ($\lambda/2$ in length to avoid reflections). The antenna is carefuly matched to the feeder (which has a characteristic impedance of 50 Ω) for two frequency bands, viz. for a bandwidth of several Mc/s around the spectral line of frequency 1420 Mc/s and for a similar band around 1360 Mc/s. The latter is the image frequency with respect to the oscillator frequency of about 1390 Mc/s, which is used in the first mixer. The degree of matching is dependent on the standing wave ratio [2]) which should approximate as closely as possible to unity. An illustration of the matching is given in *fig.* 3. The matching, and hence the output impedance of the antenna - coaxial line system changes only slightly with frequency in the bands concerned.

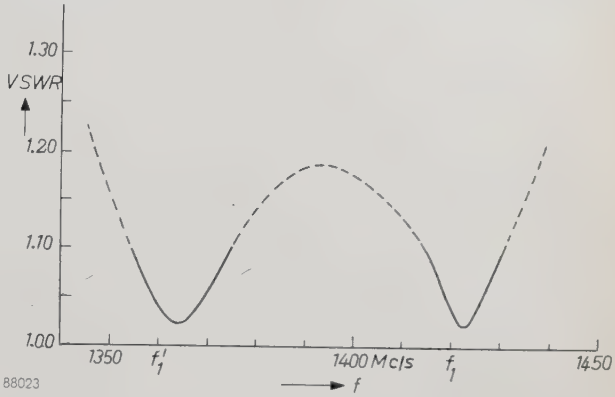

Fig. 3. Matching of antenna to 50 Ω coaxial feeder at frequencies of $f_1 = 1420$ and $f_1' = 1360$ Mc/s (wavelengths 21·12 and 22·08 cm). The degree of matching achieved is expressed by the measured *VSWR* (voltage standing wave ratio). Perfect matching occurs when $VSWR = 1$.

The weak antenna signal is fed, via a high-frequency matching transformer, to a germanium diode, which is used as the mixer valve. Thus, with no H.F. amplification and no image suppression, the H.F. section is kept relatively simple. Apart from the fact that it is not yet technically possible to achieve any appreciable gain in sensitivity by means of high-frequency amplification, its use in this receiver would in any case be undesirable. The tuned circuits involved would mean that the characteristics of the system would change too rapidly with frequency. Since the operation of the crystal mixer is closely dependent upon the impedance of its external circuit, the result would be to make it much more difficult to fulfil the requirement of equivalence between the two switching positions. This applies not only to the impedance at the signal, image and intermediate frequencies, but also to the impedance at the oscillator frequency and its multiples, and to that at frequencies which differ from these multiples by an amount equal to the intermediate frequency. The latter frequencies are also mixed, resulting in the intermediate frequency.

If the oscillator signal is very strong with respect to the input signal (as it is in this case) the current through the diode is determined exclusively by the oscillator signal. For a sinusoidal voltage across the diode, the current through it consists of unidirectional pulses. The diode current thus contains higher harmonics at multiples of the oscillator frequency. Only if the external circuit constitutes a short-circuit to all these frequencies can the voltage on the diode be purely sinusoidal. If, however, the impedance of the external circuit is infinite to one of the harmonics, the current component concerned must be zero. This means that the operation of the diode depends upon the behaviour of the external circuit at multiples of the oscillator frequency. In our receiver, the two oscillator frequencies remain constant when the entire tuning is changed. Consequently, the matching to the oscillator frequency is less important than to the signal and image frequencies, since a deviation from the requirement of equilvalent switching positions does not depend upon the tuning of the receiver.

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[2]) See J. C. van den Hoogenband and J. Stolk, Reflection and impedance measurements by means of a long transmission line, Philips tech. Rev. **16**, 309-320, 1954/55 (No.11).

The operation of the diode as a mixer stage may be likened to that of a linear network, except that in this case the input and output frequencies are different. The parameters of this network are determined by the form of the current pulse through the diode; they thus depend upon the amplitude of the oscillator signal and upon the impedance of the external circuit to multiples of the oscillator frequency. Regarding the mixer stage as a linear network, it is evident that its response, particularly at signal and image frequencies but also at the other higher frequencies mentioned above depends on the impedances connected to it.

During the switching process the high frequency section itself remains unchanged, but the frequencies in which we are interested do change. Thus, the output impedance of the H.F. section must be as independent of frequency as possible, in order to prevent the switching from affecting the circuit. Selective amplification and image suppression demand tuned circuits, which cannot easily be given a flat band-pass response. That is why neither is employed in this stage of the receiver.
A constant output impedance of the antenna - coaxial feeder system is also required for the region around the image frequency. For this reason the antenna feeder must be well matched to both the signal and the image frequency. An incidental advantage of passing the image frequency is that it facilitates the calibration of the receiver. We shall return to this point later.

It will now be clear why the switching must be effected in the first mixer stage. The equivalence requirement could certainly be fulfilled by the high-frequency section alone, but if the switching were to take place at a later stage, the preceding section of the receiver would have to have a flat band-pass response which is very difficult to achieve in practice. What is more, the effect of the inherent noise in the first amplifying stages would be much greater.

First oscillator

The antenna signal is combined in the first mixer with each of the oscillator signals in turn which are fed to the germanium diode via a capacitative coupling. The oscillator signals of about 1390 Mc/s are derived by frequency multiplication from two crystal oscillators working at 6431 and 6436 kc/s respectively. These frequencies can be varied a few kc/s by means of a variable capacitor connected in series to each crystal. The nominal values are adjusted as accurately as possible with the aid of a frequency calibration circuit.

The output voltages of both oscillators are fed to two separate two-stage amplifiers, which form part of an electronic switch. The amplifiers are alternately cut off by a negative voltage of about 50 V applied to their control grids from a multivibrator which is synchronized by a switching voltage of 400 c/s generated in a special oscillator. Two stages of amplification were found to be necessary in order to avoid capacitative coupling.

The electronic switch is followed by a frequency multiplier consisting of several stages including, inter alia, a type E 03/20 double tetrode. The total multiplication is ×216. The last multiplier stage is a trebler, which works with a disc-seal triode, type EC 55, tuned by a coaxial quarter-wave filter. This stage, together with the first mixer and a part of the first I.F. stage which follows it, is hermetically sealed in a container immediately behind the parabolic reflector. In order to lose as little as possible of the weak antenna signal to the oscillator, the coupling between oscillator and diode is very loose. For this reason it is necessary to raise the available power of the oscillator to as much as 50 or even 100 mW, although only 1mW is needed for the crystal mixer.

The signals finally delivered by the oscillators to the crystal have frequencies, f_0' and f_0'', of 1389.196 and 1390.276 Mc/s respectively, which differ by $\Delta f = 1080$ kc/s; these signals are switched to and fro at a frequency of 400 c/s.

Characteristics of the mixer circuit

The properties of the reciever may be characterized by two important quantities, viz. the conversion loss L and the noise temperature t of the mixer stage. The conversion loss is the ratio of the available antenna signal power at the signal frequency to the corresponding I.F. signal power at the output of the mixer. The value of L with a good crystal should be 5 dB. For reasons already mentioned, the conversion loss is considerably dependent on the frequency characteristics of the external circuit.

The noise temperature, which is really only a temperature ratio, expresses the relationship between the noise power originating from the mixer at an antenna temperature of 300 °K and the noise power of a resistance having the same value as the impedance of the mixer output at the intermediate frequency concerned, both being measured at the I.F. input, i.e. behind the mixer stage. The noise inherent in the mixer will always be greater than in an ordinary resistance, but with a good mixer crystal the value of t is found to be less than 1·5.

Like L, t depends upon the impedances in the equivalent circuit of the mixer stage, i.e. upon the external circuit, and also upon the output voltage of the oscillator circuit. In principle, therefore, there might be slight differences in L and t (and thus in the receiver characteristics) between the two frequencies which are switched to and fro, resulting in the appearance at the receiver output of a 400 c/s voltage component that is not due to interstellar radiation. It is therefore important to make the oscillator output signal exactly equal in both switching positions. (The use of two I.F. channels reduces somewhat the adverse effect of such a difference, as the extra signals are generally in phase in both channels and are accordingly eliminated. Nevertheless, this in itself is not enough.)

To achieve equal oscillator output a servo system is employed. The appearance of a 400 c/s A.C. component in the diode current is used as an indicator of any difference in the amplitude of the oscillator output voltage. This component is amplified, rectified in a synchronous detector circuit, and finally converted into a 50 c/s alternating voltage which drives a 50 c/s servo motor. (In principle, a 400 c/s servo motor might also have been used; present-day servo-amplifiers are available for this and other frequencies). The servo motor regulates the position of a stub in the last (coaxial) circuit of the frequency multiplier relating to the first oscillator. The time constant of this regulation is fairly large, which reduces circuit sensitivity to interfering voltages of short duration. The servo system keeps the oscillator output power in both positions equal within approximately 0.1%, which is amply sufficient at the present sensitivity of the receiver.

A further characteristic of receiver noise is the noise factor. With reference to the concept of "antenna temperature", discussed in I, the noise factor can be simply defined as follows. With an ideal receiver, i.e. one which itself produces no noise, a change in antenna temperature from room temperature T_0 to twice the value, $2T_0$, would result in twice the noise power at the output. With a non-ideal receiver the noise power at the output increases by a factor of only $(1 + {}^1/_N)$, in which N (>1) represents the noise factor of the receiver. The receiver behaves as if there appeared on the input terminals not merely the signal from the antenna but also a noise power equivalent to a temperature of $(N—1)\ T_0$. The cause is to be found in the receiver itself. The extra noise is called the inherent noise of the receiver, and it arises chiefly in the first mixer and in the first I.F. stage which follows it. Of course, noise is produced in the other stages too, but it is neglible compared with the greatly amplified noise of the first stages.

If N_{IF} is the noise factor of the I.F. amplifier alone, the noise factor of the whole receiver is

$$N = L(t + N_{\mathrm{IF}} - 1).$$

Since the noise temperature t for a good mixer crystal is at the most 1.5, it follows that N_{IF} is an important factor. The I.F. input must accordingly produce a minimum of noise. In the receiver under discussion, an N_{IF} of about 1.5 was obtained at an intermediate frequency of 30 Mc/s, which gives a noise factor of $N \approx 6$ for reception at 1420 Mc/s alone. However, the noise factor is also affected by the use of image suppression. In our case it is not used, and the total noise factor, as defined above is only half as large, i.e. $N \approx 3$, noise being received at both the signal and image frequencies instead of at the signal frequency alone. The manner in which the noise factor is measured will be discussed later.

The first I.F. section

As shown by the block diagram in fig. 2, the signal of about 30 Mc/s from the first mixer is fed to the first I.F. amplifier, which has a bandwidth of approximately 5 Mc/s. The reason for choosing such a large bandwidth is that the frequencies at which the measurement is made vary by several Mc/s in the course of traversing the line profile. As stated above, the input of this amplifier is designed to produce the minimum of inherent noise, with a view to obtaining the lowest possible noise factor. For this purpose a "cascode" circuit is used, the principles of which will not be dealt with here [3]). The amplifier consists of two sections: the cascode circuit, together with one of the amplifier stages, is mounted directly behind the parabolic reflector in the hermetically sealed container already mentioned, and the next two conventional amplifier stages are situated in the observation cabin with the rest of the receiver. The sections are joined by a screened coaxial cable.

To the second section of the amplifier a suppressor circuit is connected, the purpose of which is to ensure that the switching process itself has no effect on the measurements. In spite of all the measures mentioned, it was impossible to satisfy this condition without using a suppressor, for the following reason. The process of switching produces voltage pulses which manifest themselves throughout a wide band of frequencies and which, owing to the first mixer, also appear with appreciable amplitude in the I.F. band at 30 Mc/s. This would present no problem if switching from the second to the first position produced exactly the same waveform as switching from the first to the second. In that case the frequency of the switching pulses would be 800 c/s, to which the synchronous detector at the end of the receiver is insensitive. In actual fact, however, the pulses are dissimilar, with the result that a 400 c/s component appears at the output of the I.F. amplifier, causing spurious deflection of the recorder. As the amount of this spurious deflection depends on the tuning frequency and on the temperature, it is not constant and is therefore very troublesome during measurements.

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[3]) H. Wallman, A. B. MacNee and C. P. Gadsden, Proc. Inst. Radio Engrs. **36,** 700-708, 1948. See also H. H. van Abbe, B. G. Dammers and A. W. G. Uitjens, Noise of the cascode amplifier, Electronic Appl. Bull. **14,** 141-150, 1953.

The suppressor cuts out the I.F. amplifier just before the moment at which the oscillators are switched over, and switches it on again immediately after. This, ofcourse, introducess witching pulses from the suppressor, but the form of each pulse is now identical. The suppressor consists of a mono-stable "puller" circuit [4]), driven by pulses with a frequency of 800 c/s derived from the 400 c/s switching voltage appropriately shifted in phase. The second section of the first I.F. amplifier remains switched off on each occasion for approximately 50 μsec.

Nevertheless, a 400 c/s component not originating from the antenna might still appear if the spacing of the cut-out pulses is unequal, owing to incorrect adjustment of the 400 c/s driving voltage in the suppressor. However, this fault can easily be corrected, and as it is independent of frequency and thus identical and in phase in both I.F. channels, it does not in the end affect the measurement.

The remaining I.F. stages

The 30 Mc/s signal is fed to a second mixer stage, containing a standard ECH81 heptode, in which it is mixed with the signal from a second oscillator. The frequency of this oscillator is varied continuously between 34 and 39 Mc/s by means of a small high stability tuning capacitor driven via a worm gear by a synchronous motor. In this way the whole profile of the spectral line is traversed.

The frequency of the difference signal leaving the mixer varies between 4 and 9 Mc/s. After further amplification, the signal divides into the two separate I.F. channels via two amplifiers in parallel operating at 4·96 and 6·04 Mc/s respectively. The output signals from these amplifiers are each combined in third mixers with the 5·50 Mc/s signal from a common crystal oscillator. Thus, the resultant difference signals in both channels have the same frequency, viz. 0·54 Mc/s. Each amplifier attenuates the image frequencies by more than 30 dB.

The two 0·54 Mc/s signals are then each fed to 3-stage amplifiers, each having a frequency characteristic determined by three identical resonant elements. The bandwidth of both is 35 kc/s. As already mentioned, these amplifiers determine the bandwidth of the channels; all other amplifiers have much larger bandwidths.

It is not necessary, nor is it desirable, to have flat band-pass characteristics. The present characteristics are approximately Gaussian in form which facilitates frequency adjustment. This is an important point, because the tuning to 0·54 Mc/s must be exact. Actually, the tuning is so stable that it scarcely needs any readjustment. Diode detection follows the last stage in both channels. The D.C. output is taken from the diode load resistor and fed via a low-pass filter to the amplifier for automatic gain control. The latter is necessary because the amplitude of the 400 c/s square wave voltage naturally depends upon the gain of the I.F. amplifier For reasons of simplicity, it is not actually the gain that is kept constant but the average output voltage, which can be tapped off directly from the detector output. Strictly speaking the noise output voltage (i.e. in absence of signal) rather than the average output voltage should be kept constant in order to achieve the desired constant gain. However the errors due to use of the average output voltage are negligible and corrections are unnecessary

In our reciever the A.G.C. works on the last mixer valve and on the amplifier valves preceding it. Its effective time constant is approximately 0·2 sec, so that the 400 c/s component is not affected.

The synchronous detector circuit

The low-frequency alternating voltage across the two load resistors consists of different components. Apart from the 400 c/s square-wave voltage to be measured, it includes the suppressor pulses, which have a repetition frequency of 800 c/s, and the inherent noise of the receiver, in which frequencies from 0 to 35 kc/s are represented. The two 400 c/s components are equal and in antiphase, the 800 c/s pulses are also equal but in phase, while the noise voltages are mutually uncorrelated.

The synchronous voltmeter measures the fundamental of the 400 c/s square-wave voltage and the amplitude of that part of the noise spectrum which is immediately adjacent to this frequency. Of the entire spectrum only a very narrow band, about $^1/_{200}$ c/s around the 400 c/s switching frequency, passes beyond the diode detectors and is recorded. The detector circuit which achieves this high selectivity functions as follows.

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[4]) B. Chance, Wave forms, Radiation Laboratory Series No. 19, McGraw Hill, New York 1949, page 181.

The two 400 c/s components are added in a push-pull input circuit and the two in-phase 800 c/s components thus eliminated. This circuit is followed by a selective amplifier with a bandwidth of about 100 c/s. Its function is to attenuate the odd harmonics of 400 c/s (to which the synchronous detector is also sensitive) sufficiently to make the total sensitivity to these frequencies neglible.

The synchronous detector itself (*fig.* 4) is really a mixer stage, to which is applied not only the 400 c/s signal voltage, but also a relatively large oscillator-voltage with exactly the same frequency. The difference frequency is therefore zero, that is to say the result is a direct voltage proportional to the amplitude of the signal — and to the cosine of the phase angle between the two 400 c/s voltages. The detector proper is a bridge circuit, each arm consisting of a vacuum diode and a $100\ \text{k}\Omega$ wire resistor in series [5]). The signal and the oscillator voltages are symmetrically applied to opposite diagonals and the D.C. output voltage appears between the centres of two resistance arms across each push-pull cathode follower output (see fig. 4). As the oscillator voltage is large compared to the signal voltage one half of the diode bridge periodically becomes conducting while the other half presents an infinite resistance. The resistance of the conducting branch is determined mainly by the wire resistors, so that differences in the diode characteristics have very little effect. An examination of the current flow in the bridge shows that, if the two voltages are in phase, a D.C. voltage appears across the output.

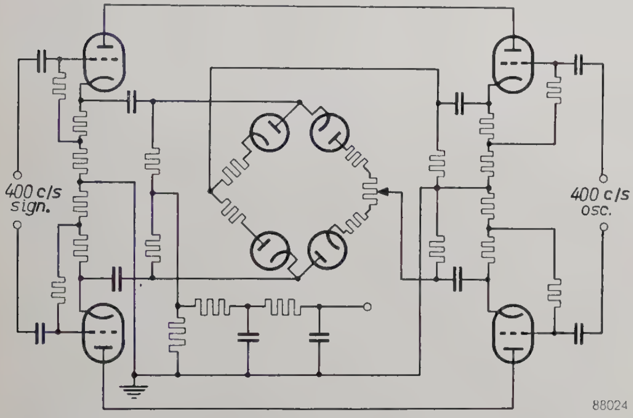

Fig. 4. Simplified circuit of the synchronous detector.

The fundamental of the 400 c/s square-wave voltage having thus been filtered out, the output D.C. voltage is fed to a low-pass filter with a bandwidth of about 0·005 c/s, which determines the bandwidth of the voltmeter. The filter consists of a double RC network, which acts as an almost critically damped filter. The time constant of the filter is inversely proportional to the bandwidth, and may therefore be used instead of the bandwidth to describe the characteristics of the filter. The time constant, as defined by Nechleba [6]), is 54 secs in the present case.

The filter is followed by a D.C. amplifier which is used purely and simply as an impedance transformer: the amplifier input does not load the low-pass filter. The output impedance of the amplifier [7]) is about $1\ \Omega$, which is lower than the input impedance of the recording meter.

The narrower the bandwidth of the circuit the less will be passed of the interfering noise spectrum. However, this also necessitates slower variation of the second oscillator, because the tuning rate, expressed in I.F. bandwidth per second, determines the smallest admissible bandwidth of the voltmeter. It can be shown that at a tuning rate of n bandwidths per second the cut-off frequency of the low-pass filter must lie at about n c/s.

If the intensity of the spectral line is measured as a function of the frequency with a band-pass filter of bandwidth Δf (in our case 35 kc/s) the measured intensity will always be an average of the "true" intensity distribution over this bandwidth. This implies that details in the "true" intensity distribution, whose width is narrower than the bandwidth, are not resolved, or are observed only in an attenuated and distorted form.

If we analyse this further, assuming the bandpass curve to be sinusoidal (*fig.* 5), we find that, at a tuning rate of n bandwidths per second, hardly any frequencies equal to or greater than n can occur in the output voltage behind the synchronous detector. The low-pass filter does not, therefore, require a bandwidth larger than n. However, if this bandwidth were much narrower, the highest frequencies in the output voltage would not be passed, so that the profile observed behind the filter would look as if the bandwidth of the I.F. band-pass filter were larger. In our receiver, the bandwidth of the low-pass filter has been chosen such that its 3dB cut-off frequency (i.e. the frequency at which the power passed has fallen to one half of its value) just coincides with the above-mentioned limit of n c/s. In this case the low-pass filter has hardly any influence upon the form of the observed line profile. The only noteworthy effect is that the profile is subject to a delay during its passage through the low-pass filter. This delay is equal to the time constant of the filter (54 sec) so that in the receiver under consideration the intensity recorded on the meter at any given moment refers to the frequency to which the receiver was tuned 54 seconds earlier.

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[5]) Cf. a discussion of the ring modulator by F. A. de Groot and P. J. den Haan in Modulators for carrier telephony, Philips tech. Rev. 7, 83-91, 1942.

[6]) F. Nechleba, Elektrotechn. Z. **74**, 98-101, 1953.

[7]) C. Morton, Electronic Eng. **25**, 312, 1953.

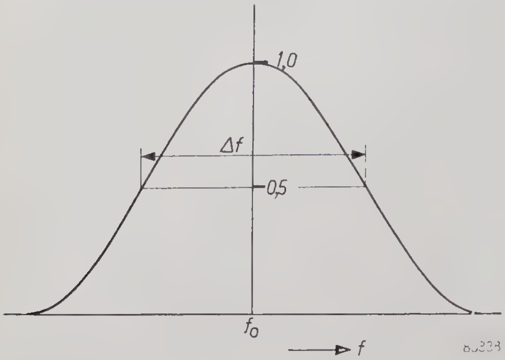

Fig. 5. Sinusoidal band-pass response of the I.F. filter.

The bandwidth of a filter with a rectangular band-pass characteristic must be considerably larger, since the intensity of the voltages with frequencies above n c/s is then no longer neglible. However, these frequencies are not very important in the measurement of the profile as they are considerably distorted with respect to the true profile. (Between n and $2n$ there is a phase shift of exactly 180° with respect to the true profile). Quite apart from these considerations, a rectangular band-pass characteristic is more difficult to achieve in practice than a sinusoidal one; for this reason the sinusoidal characteristic adopted in the present receiver is to be preferred.

The theoretical sensitivity of the receiver

By theoretical sensitivity is meant the smallest difference in antenna temperature between signal and comparison channels that is still theoretically measurable. It depends upon the method of measurement and upon the noise factor, the I.F. bandwidth and the time constant of the voltmeter. The determining factor for the sensitivity is the magnitude of the noise fluctuations during the recording of the line profile; their amplitude can be directly expressed in units of antenna temperature i.e. in degrees Kelvin. If we assume, according to accepted usage, that the smallest measurable change in antenna temperature is equal to the effective value of the noise fluctuations, then the theoretical sensitivity is equal to the theoretically calculated effective value of the noise fluctuations, expressed in degrees K.

It should be noted that the smallest measurable difference in antenna temperature between the two channels is, in reality, dependent not only upon the magnitude of the noise fluctuations but also upon the time duration of such a difference. If the antenna temperature first has a substantially constant value for a period of time that is long with respect to the time constant, and if it then quickly moves to another constant value, the smallest measurable difference will indeed be approximately equal to the effective value of the noise fluctuations. For a difference of only short duration, however, the sensitivity will be lower, because there is a greater chance that the deflection on the meter was not caused by a difference in antenna temperature but by an incidental noise-fluctuation, which may well be greater than the average fluctuation.

It will be sufficient for our purpose to consider only the effective value of the noise fluctuations. Expressed in degrees K it is given by:

$$\Delta T = 1{\cdot}03\,\frac{\pi}{\sqrt{2}}\,(N{-}1)\;T_0\,(F\Delta f\cdot\tau)^{-\frac{1}{2}},$$

where N is the noise factor for simultaneous reception at signal and image frequencies, $T_0 = 293$ °K, Δf is the bandwidth of the I.F. amplifier, τ is the time constant of the low-pass filter, and F is a factor determined by the shape of the I.F. band-pass characteristic.

A similar formula was first derived by Dicke [8]) for a receiver with one I.F. channel and a square law detector. The theoretical sensitivity of our receiver with two I.F. channels is superior by a factor $\sqrt{2}$, which may be simply explained as follows. The addition of the voltages behind the two diode detectors doubles the magnitude of the 400 c/s signal voltage and increases the noise voltage by a factor of only $\sqrt{2}$, there being no correlation between the two noise voltages. The signal-to-noise ratio is therefore $\sqrt{2}$ times larger. Taking into account that the spectral line is received only at 1420 Mc/s and not at the image frequency, 1360 Mc/s, and that a linear diode detector is used instead of a square law detector, which reduces the sensitivity by 3%, the above formula can be derived directly from that of Dicke.

In the network used, which consists of three identical resonant circuits, $F = 1{\cdot}53$, $N \approx 3$, $\Delta f \approx 40$ kc/s (as regards the fluctuations of inherent noise, it is better to use this value of the 3dB cut-off frequency than the earlier-mentioned bandwidth of 35 kc/s) and $\tau = 54$ secs. The theoretical sensitivity is thus:

$$\Delta T = 0{\cdot}73\ ^{\circ}\mathrm{K}.$$

If the theoretical sensitivity of this receiver is compared with that of a receiver in which no switching takes place, the present receiver is found to be inferior by a factor of 1·57. This factor is mainly attributable to the fact that the present receiver records the difference in antenna temperature in signal and comparison channels (with the same bandwidth), the uncorrelated noise voltages being quadratically added, whereas the ideal receiver records only the increase of antenna temperature in one channel. The sensitivity thereby deteriorates by a factor of $\sqrt{2}$; the remainder, a factor of 1·11, is concerned with the synchronous detector. The loss in sensitivity resulting from the method of frequency comparison is amply offset by the much greater stability obtained. In practice, the zero stability during recording decides the maximum sensitivity obtainable with a receiver. From the formula given for ΔT it would appear that, in principle, it is possible to achieve any sensitivity required by increasing the time constant (or the I.F. bandwidth, which is not desirable in this case, however, as it entails the loss of details in the spectrum). However, a corresponding reduction in the tuning rate is then necessary and this prolongs the measurements. The real limit is therefore set by the zero stability, which is subject to a variety of factors such as fluctuations in the supply voltages and changes in ambient temperature. The sensitivity of the receiver while it was located at Kootwijk was restricted by interference from neighbouring transmitters to a limit of about 1 °K. This means that variations in signal intensity amounting to less than 0.1% of the noise inherent in the receiver could still be recorded. The performance of the receiver in its new and quieter location at Dwingelo is not yet known. A substantial improvement in sensitivity is to be expected.

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[8]) R. H. Dicke, Rev. sci. Instr. **17**, 268-275. 1946.

Frequency and intensity calibration

The various frequencies generated in the receiver itself must be kept strictly constant if the measurement of the line profile is to be reliable. This applies particularly to the oscillator frequencies. The switching frequency need not be exactly constant, since the electronic switch and the tuned detector operate at the same frequency, derived from the same oscillator.

The accuracy required is 1 in 10^6, and the frequency scale on the recorder is therefore accurate to within 1·4 kc/s. The frequency calibration is carried out in three stages (*fig. 6*):

1) Adjustment of the two crystal oscillators in the first mixer circuit.
2) Calibration of the slowly varying frequency of the second oscillator.
3) Calibration of the pass bands of both I.F. amplifiers behind the third mixer to a width of 35 kc/s.

Kootwijk Radio Station generates standard frequencies of 1 to 10 kc/s with an accuracy of 1 in 10^7; they are derived from the frequency standard adopted by the Netherlands Post Office in the Hague the accuracy of which is better than 1 in 10^9. These frequencies come within the range of the present receiver and are therefore used as a primary standard for calibrating the secondary frequency standard in the receiver itself. The secondary standard is a 100 kc/s oscillator, whose driving crystal is contained in a thermostatted enclosure and the frequency it generates is compared on an oscilloscope with the primary calibration frequency. The accuracy of the calibration is about 1 in 10^7.

The frequencies of the oscillators in the first mixer stage should, after being trebled, be 19.308 and 19.293 Mc/s respectively. A frequency of 19.300 Mc/s is now derived from the secondary standard (by multiplying it by 193) and a voltage of this frequency is mixed with the two signals mentioned. The difference frequencies of 7 and 8 kc/s can be directly compared on an oscilloscope with the primary standard of 1 kc/s.

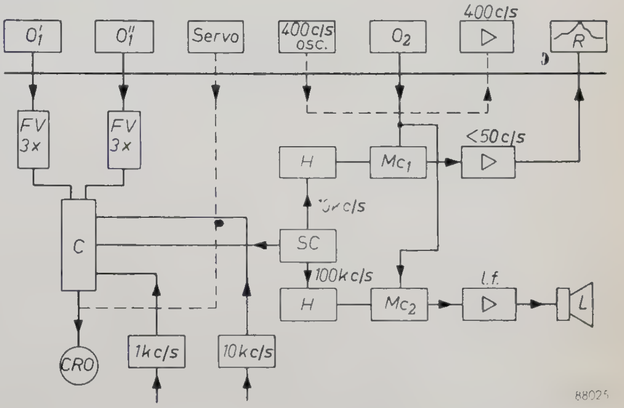

Fig. 6. Simplified block diagram of the frequency calibration circuits (full lines) and control circuits (broken lines). The units above the horizontal line are parts of the receiver proper (see fig. 2). *FV* frequency multipliers, *C* calibration panel, *KO* cathode ray oscilloscope, *SC* secondary frequency standard, *H* harmonic generators, Mc_1 and Mc_2 mixer circuits, *L* loudspeaker.

The second oscillator varies slowly in frequency, and a different method of calibration has to be adopted. The secondary standard is divided in order to obtain a frequency of 10 kc/s, and multiples are then derived from this with frequencies between 33 and 39 Mc/s. After amplification, this entire group of harmonics is mixed with a voltage having the varying oscillator-frequency, one result of which is the appearance of a difference frequency somewhere between 0 and 5 kc/s. This signal is fed to the recording meter via an L.F. amplifier with a pass band of only 0·5 to 50 c/s. Only if the difference between the oscillator frequency and one of the multiples (all are multiples of 10 kc/s) is smaller than 50 c/s will a signal be perceptible at the output of the receiver. After rectification, this signal actuates a relay which short-circuits for a moment the recording meter and causes the pen to fall towards zero. Thus, every time the oscillator frequency

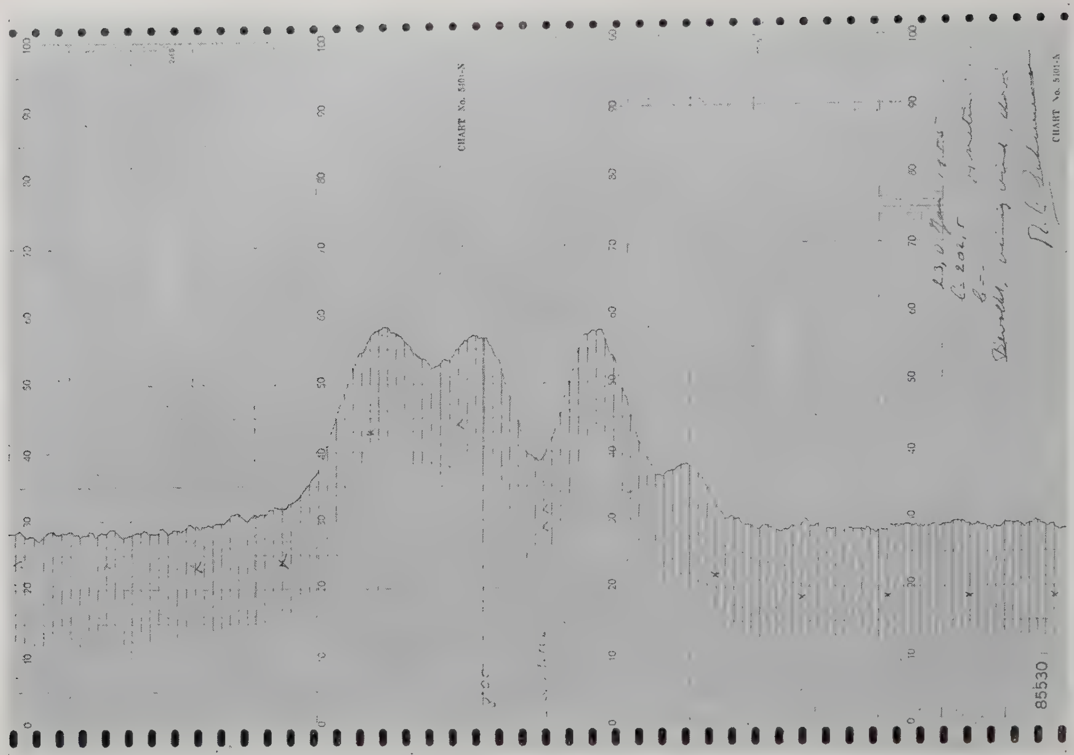

Fig. 7. Example of a recording. The vertical marks are at frequency intervals of 10 kc/s. Multiples of 100 kc/s are marked with a cross. This recording was made on the night of 23rd January 1955, for radiation from a point in the sky given by $l = 202{\cdot}5°$, $b = 0$. At this bearing several maxima are observed (implying spiral distribution of the interstellar hydrogen).

passes a multiple of 10 kc/s, a mark is automatically made on the strip chart (*fig.* 7).

A similar mark is made at multiples of 100 kc/s or 1 Mc/s is passed (with the aid of specially generated multiples of these frequencies), but they are also made audible via a loudspeaker. The observer can thus indicate on the recording what marks correspond to what multiples.

The band-pass characteristics of the two I.F. amplifiers following the third mixer are automatically recorded in the receiver with the aid of a multiple of the 100 kc/s secondary standard which lies in the proximity of 30 Mc/s and which can be fed to the first I.F. amplifier. The pass-band has proved to be very constant and need be checked only once a week at the most.

Intensity calibrations are only relative measurements carried out during recordings over several months. A daily measurement is made of one part of a line profile received from a fixed point in the sky. This profile is taken as the standard with which the intensity of all other profiles measured is compared. To determine the absolute intensity scale it is therefore only necessary to calibrate now and then the intensity of the standard profile.

Direct absolute calibration with an artificially obtained spectral line is not possible, as no noise source is available that emits radiation in a spectral line and whose intensity can be determined independently of a radio wave measurement by using a well-known physical relation (in the way, for example, the radiation from a black body can be ascertained by measuring the temperature of the body). Absolute calibration therefore has to be done indirectly by calibrating the various receiver-constants concerned, such as the noise factor, the rectification characteristic of the two diode detectors and the sensitivity of the 400 c/s voltmeter. The diode characteristic is easily determined by connecting a noise diode instead of the first mixer to the 30 Mc/s I.F. amplifier. The 400 c/s voltmeter can be calibrated quite simply by means of a known 400 c/s voltage.

Several factors must be taken into account, however, when measuring the noise factor, Since no image-suppression is employed, the noise factor ascertained with a given noise source always holds for simultaneous reception on signal and image frequencies. Now for measuring the spectral line, we are concerned only with reception at 1420 Mc/s, so that the sensitivity ratio at the two frequencies must be found in another way. The method used here is to make the sensitivity at both frequencies approximately equal by increasing the bandwidth of the mixer stage. It remains to be seen how sensitive the receiver will be to the odd multiples that arise during mixing. (A short-circuit is used for the even multiples in the mixer stage and in the antenna). It is to be expected that the noise power received at these frequencies will amount to less than 10% of that received at the signal frequency. Since these properties have not yet been determined in detail, the antenna temperature scale in use at present is only temporary and may conceivably involve an error of 20%.

For determining the noise factor earth radiation is used as the noise source. For this purpose the antenna was first directed towards the zenith and then upon a pine-wood situated on a slope to the north-east of the antenna site. The continuous radiation received was found to increase by 50%. If we assume that the earth emits radiation like a black body with a temperature of 300 °K, while the antenna temperature in the other position (zenith) amounts to only a few °K, say 5 °K, then the difference in antenna temperature between the two positions will be 295 °K, provided that the pine-wood radiation fills the whole antenna pattern. If it does not, the temperature difference will be smaller, which means that the real noise-factor will be lower than the value found by measurement. This method, therefore, allows only the maximum value of the noise factor to be determined, which in our case will not differ much from the real value. Thus, for simultaneous reception at 1420 and 1360 Mc/s, we find $N_{\max} = 3.0$; for reception at 1420 Mc/s alone, assuming equal sensitivity, the noise factor is then $N'_{\max} = 6.0$.

From the temporary intensity scale calibrated in this way the highest value of the spectral line temperature measured was 118 °K.

Power supply

The entire receiver is fed from the electric light mains. The mains voltage is kept constant to 0·5% by a voltage stabilizer of 2 kVA. The high tension is taken from a number of electronically stabilized supply units of the same design, incorporating a neon tube type 85A1 or 85A2. Further stabilization is unnecessary in this receiver since, with the method of switching used and the automatic gain ocntrol, the operation of the measuring circuit is largely independent of the supply voltages.

The receiver discussed in this article is the result of several years of gradual development, periods of improvement having alternated with prolonged periods of continuous measurement. In the form described, the receiver has been in almost uninterrupted operation from August 1954 to August 1955. As regards sensitivity it has come up to all expectations. Zero stability, too, is excellent, so that reduction of the observations to give the line profile presents no special problems. Although the receiver contains more than 200 valves, scarcely any trouble has occurred during operation. In this respect it may be noted that the receiver is left switched on during intervals between recordings and that the valves in general operate far below their maximum anode dissipation. Another favourable factor is the stabilization of the supply voltage. All valves are inspected once every six months, when 10 to 20 valves usually have to be replaced.

The equipment is operated mainly by students of Leiden University, most of whom are not specialists in electronics. After a day or two of instruction, however, they can safely be left to carry out the measurements without supervision.

Summary. This article deals with the design of a receiver for radio waves with a frequency of 1420 Mc/s which are emitted by interstellar hydrogen. In order to be able to distinguish the very weak signal from the noise inherent in the receiver, a method of frequency comparison is employed. The receiver is switched to and fro at a rate of 400 times per second between two frequencies, one lying on the spectral line and the other adjacent to it. The spectral line is traversed by slowly varying both frequencies simultaneously while maintaining a constant difference of 1080 kc/s between them. Two separate I.F. channels are used in order to keep the receiver tuned to the spectral line all the time (instead of only half the time, as would be the case if only one I.F. chamnel were used). Each channel has its own comparison frequency on opposite sides of the measuring frequency, and each is tuned to the line in turn. The receiver must satisfy two primary requirements: its characteristics must be identical in both switching positions, and the process of switching must in no way affect the measurement. To fulfil the first condition, the high-frequency section is of simple design and is carefully matched to the first mixer stage at the signal and image frequencies. The oscillator frequency of this stage springs to and fro at a rate of 400 c/s between two settings 1080 kc/s apart. The output voltage of the oscillator circuit is made identical in both positions by means of a servo system. The second condition is fulfilled with the aid of a special suppressor circuit, which cuts off the receiver for a moment during the switchover. Some stages of amplification are then followed by the second mixer, the oscillator frequency of which is slowly varied for the purpose of traversing the spectral line. Division now takes place into the two I.F. channels mentioned, which are each provided with automatic gain control. These two channels determine the ultimate bandwidth of the receiver. After further mixing and subsequent detection the two signals are fed to a push-pull amplifier, where they are added together and where unwanted voltages are, as far as possible, eliminated. The fundamental frequency of the 400 c/s square-wave voltage, which is a measure of the intensity of the interstellar radiation, is rectified in a tuned detector circuit and, after emerging from a low-pass filter of only $^1/_{200}$ c/s bandwidth, it is fed to a strip-chart recorder. The frequencies generated in the receiver and the bandpass characteristics of the I.F. filters, which must be identical and constant in both channels, can be verified and adjusted by means of various semi-automatic calibration circuits.
The article finally deals briefly with the noise characteristics of the high-frequency section and of the first mixer, and discusses the factors that determine the sensitivity of the receiver. The sensitivity achieved is better than 0.1% of the inherent noise. In terms of antenna temperature, this means that it is possible to detect temperature differences down to about 1 °K.

SCIENCE GOES TO THE FAIR

by G. REMEDI.

53:79

The post-war growth of Philips has been accompanied by a continual increase in the personnel of the Research Laboratories. From 1946 to 1956 the staff increased from about 700 to about 1300. To provide accomodation for these growing numbers the laboratory has been repeatedly extended. The fortieth anniversary of the laboratory in June 1955 saw the completion of a new wing. These events were celebrated at a specially organized fair which, on account of its scientific nature, deserves some description in this review. The author of this article is one of the leading figures among the organizers, writing under a nom-de-plume.

Scientific research has been compared to a game and even in an industrial laboratory daily activities often have something of this character. One of the most interesting aspects of this game with Nature is that as it progresses both the strategy and the equipment must be continually revised and adapted to the changing situation.

The impact of this element of play on the staff of the Research Laboratory found another outlet on the 2nd of July 1955, when the entire laboratory personnel and many former members, their wives and fiancées, celebrated the fortieth anniversary of the laboratory and the official opening of a new wing (*fig. 1*).

For this monster celebration (2800 people were present) a large hall in Eindhoven was converted into a Brabant village square on which a "scientificfair" was set up. The title photograph, taken during its construction, gives an impression of the fair. The side-shows were prepared by the various groups of the laboratory, each contribution bearing some reference to the daily work and equipment of the relevant group. Relays of laboratory personnel took it in turns to run the fair so that everyone had an opportunity to enjoy it as visitors. In addition to the sideshows there was a general programme with dancing, competitions, a ballet and an opera with well-known arias but a new libretto appropriate to the occasion.

A "glass" dance-floor was made by a method commonly used for the construction of film scenery. Strong paper is stuck on a flat foundation and painted with coloured motifs. A few layers of water-glass, painted-on, provide both an ideal surface and excellent durability (see title photograph). The durability of the paper and other materials was first tested on a small scale in the laboratory with a thoroughness typical of the enthusiasm with which all the preparations were tackled. Advantages of a dance-floor of this type are its glossy and colourful appearance which breaks up the monotony of a large expanse of floor, its adaptability as regards shape and extent and the low price. The life of such a floor can be easily extended, if need be, simply by applying a new layer of water-glass.

The musical background for the event was provided by a professional dance-orchestra, a barrel organ and the laboratory's own band and, in addition, an "electronic orchestra". The instruments of the latter, designed and played by members of the acoustical group, each have their own amplifier and loudspeaker, together covering the whole audible frequency range and giving an output loud enough to be heard everywhere above the din of the fair. The melody section was made up of four non-polychromatic instruments. The soprano and tenor parts were played on two multivibrators, the pitch being controlled by variable resistors. The alto part was supplied by an A.F. generator, played by means of the frequency control knob whilst a low-frequency *RC*-oscillator was used for the bass. The rhythm section included kettle drums in the form of two large capacitor-microphones, the diaphragms (8″ and 6″) being played with the fingers like drum-skins. In addition, there was a pulsed noise generator that acted as cymbals, and a device imitating a bass drum. In the circuit designed for this drum, a capacitor was discharged across an *LC* circuit; the amplifier was provided with positive feedback to give the required resonance (*fig. 2*). An electrical clavichord completed the ensemble. Its strings were stretched tightly across a metal plate, forming a series of capacitors, so that, when a string vibrates, an electrical signal is produced. Just in the same way as on a normal clavichord it was possible to play polyphonically and even to reproduce the characteristic vibrato of a clavichord. The volume could be controlled electrically by a pedal as well as by the touch on the keys. Visitors were particularly fascinated by the apparent ease with which the members of the orchestra produced so much volume and richness of tone.

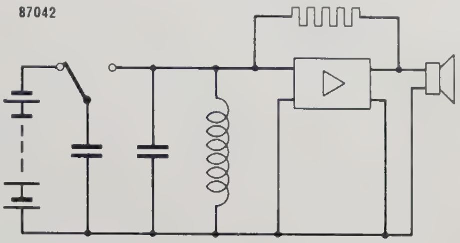

Fig. 2. Circuit diagram of the electronic bass drum.

Points were awarded in both the general competitions and in the competitive side-shows. Prizes were awarded according to the total number of points which each participant had collected. With the help

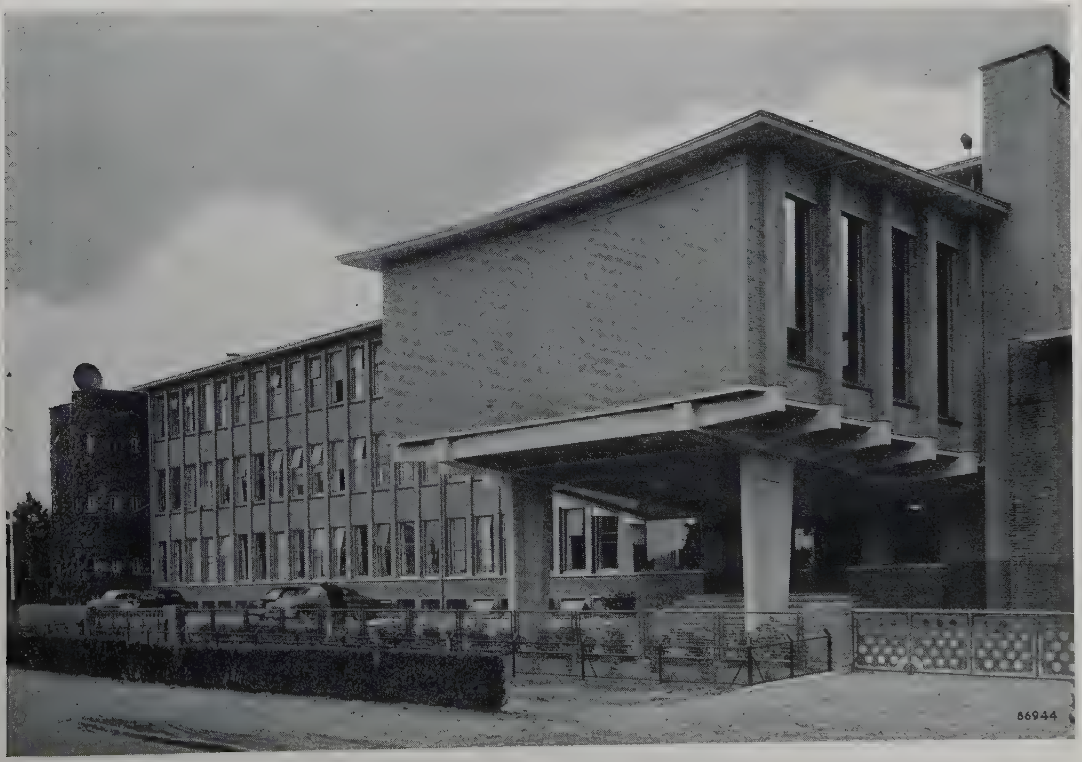

Fig. 1. The new wing of the Research Laboratories at Eindhoven, the official opening of which on 2 July 1955 was celebrated by the scientific fair described in these pages.

of the statistical department the system of allocating points was so arranged that, at the higher end of the scale, the chance of two or more persons ending up with the same number of points was very small.

A number of original contrivances were prepared for the main competitions. Among them were the "Hariguars", 3-wheeled racing cars equipped with independent front wheel hand drive with a different transmission ratio for left wheel and right wheel, and with a swivelling wheel at the back. To drive in a straight line in this vehicle requires considerable skill.

During the rehearsal of the opera, it was found that a well-nigh perfect performance could be ensured by making use of the methods of modern electronics. Because of the poor acoustics and the size of the hall it was necessary to reproduce the dialogue and the arias through amplifiers to make them intellegible. Direct reproduction has the disadvantage that the players are more less or bound to the microphone. This was avoided by recording the whole opera (with the exception of the overture which was played *au naturel* by the wind-orchestra) on magnetic tape, the players (without microphones) merely singing and acting synchronously with the tape during the actual performance. It is remarkable how word-perfect the performers can be by this technique!

The stalls at the fair were of two types: those with a competitive element where points could be won towards the prizes and those sideshows which were merely for entertainment. As already mentioned, the stalls were run by members of the laboratory staff in rotation; each person spent about one hour on duty and in this way more than 1000 fair-goers took an active part in the success of the fair. In all there were about 60 stalls, many of which were completely novel in character. To give some idea of the nature of these scientific playthings, technical details of some of them are given below.

Rubber membrane races (fig. 3). Races between miniature motorcycles on the taut rubber membrane used in the laboratory as an analogue for problems of electronic tubes and wave guides [1]). One end of the rubber membrane is made to vibrate vertically by means of a motor-driven eccentric so that a standing wave is set up on the membrane. The positions of nodes and antinodes depend upon the frequency, which is controlled by a "frequency twist-grip" on a pair of motorcycle handlebars. A miniature motorcycle is placed on the membrane and is so arranged that it moves forward when it is in a place where the membrane is moving. The forward speed is greatest at the position of the antinodes whilst the motorcycle remains still at the nodes. In order to reach the finishing line as quickly as possible it is necessary to vary the standing wave pattern by altering the frequency whenever the motorcycle reaches a node.

Child's play, or blowing up balloons with liquid air. Air taken from the room is liquified by means of a gas refrigerating machine [2]) and flows continuously into a Dewar-flask. The liquid is pumped out of the flask and up to the desired level by means of a vapour bubble pump. A deflated balloon fitted with a plastic funnel is then held under the stream of liquid air so that the liquid air runs into the balloon, after which the funnel tube is closed tightly with a stopper. The liquid air in the balloon vaporizes and inflates the balloon; however, the rubber of the balloon wetted by the liquid air is hard and cannot stretch. Contestants must then try to warm up these cold hard regions with a jet of air quickly enough to prevent the balloon bursting. The larger the balloon, the more points gained.

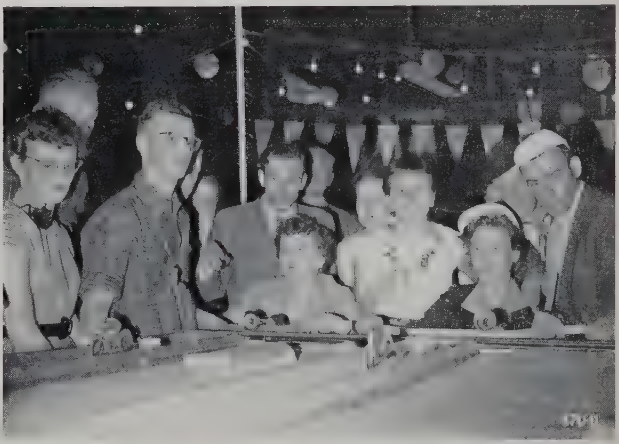

Fig. 3. The "Rubber membrane races". On the extreme left and near the centre of the photograph can be seen two miniature motorcyclists on the taut rubber membrane. The two people seated behind the membrane each grasp the frequency twist-grip on a pair of motor-cycle handle-bars. By controlling the frequency, the motorcyclist can be made to move forwards.

Philirose. This is a shop window dummy with a glass head, the glass being made locally conducting by a coating of tin oxide (*fig.* 4). Current is supplied to these regions through silver strips which are connected to a variable transformer via connecting terminals concealed in the hair. The head is coated with a paint containing Ag_2HgI_4. A handle is provided which via a train of gears rotates a governor; the latter operates the variable transformer and thus the current through the conducting glass. The heat produced makes the dummy's cheeks blush by changing the colour of Ag_2HgI_4 in the paint, which is bright yellow below 50° C and orange-red above. In addition, the brown-painted pupils dilate into a red iris by a change in the colour of Cu_2HgI_4 from bright red to dark brown at 70° C. If the handle is turned too rapidly the governor switches the current off; the handle has therefore to be turned at a particular rate to get the quickest results.

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[1]) K. S. Knol and G. Diemer, A model for studying electromagnetic waves in rectangular wave guides. Philips tech. Rev. **11**, 156-163, 1949/50. Further articles on the application of the rubber membrane can be found in Philips tech. Rev. **2**, 338-345, 1937 and **14**, 336-344, 1952/1953.

[2]) J. W. L. Köhler and C. O. Jonkers, Philips tech. Rev. **16**, 69-78 and 105-155, 1954.

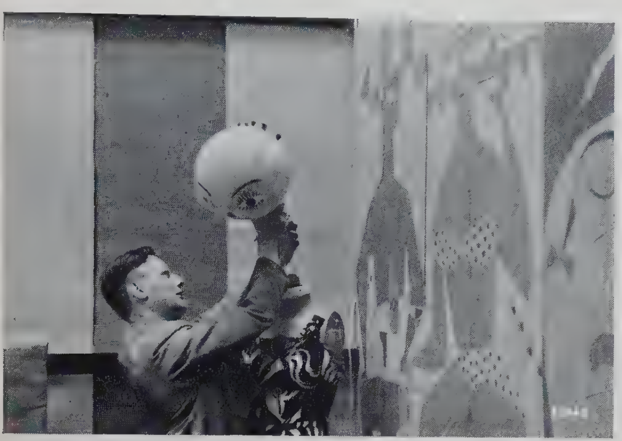

Fig. 4. Mounting the head of "Philirose". As the hair is still missing, the connecting terminals for the current supply wires can be seen on the head.

Potentiometer. Quite unlike the instrument normally associated with this word, this contrivance pretends to be the substantiation of a literal interpretation of the word — a test of strength. It consists of a bicycle on a stand, a copper disc acting as the rear wheel, which can be braked by an electromagnet (eddy current brake). This bicycle is actually used in the laboratory for physiological tests. Brute force, however, is of little use on this "potentiometer". Instead, patience and attention are required. Against the copper disc runs a bicycle dynamo which generates an A.C. voltage having a frequency proportional to the speed of the disc. This A.C. voltage is supplied via an amplifier to one of the windings of an induction motor. The other winding, perpendicular to the first, is supplied from the mains. The motor can turn only if the frequency of the voltage supplied by the bicycle dynamo is 50 c/s; and the direction in which it then turns depends on whether this voltage is leading or lagging in phase with respect to the mains voltage. By means of a reduction gear, the motor drives an endless cord on which is fastened a miniature racing cyclist who cycles over a course in a painted mountain landscape. It is quite feasible to keep the frequency practically constant at 50 c/s with pedalling but it is impossible to control the phase so that the cyclist moves mainly forwards. For this reason a slip coupling is incorporated in the transmission from motor to cord which ensures that cycling forwards is easier than cycling backwards; with this modification there is a fair chance of making the cyclist reach the finishing line within 1½ minutes. When the cyclist comes to the foot of a mountain the brake magnet is energized and the eddy current brake comes into action. As the slope becomes steeper the excitation becomes stronger and with a little imagination the competitor has the feeling of struggling up a mountain.

Knijp van Jut, or Test your Clench. Two players sit at opposite sides of a table each operating a hand dynamo [3]. The hand dynamo supplies an A.C. voltage whose frequency is proportional to its speed of rotation. This A.C. voltage is fed to a low-pass filter and is then rectified (*fig. 5*). Since the amplitude of the A.C. voltage also increases with the frequency, a maximum D.C. voltage is obtained if the hand dynamo is operated at such a rhythm that the frequency lies just below the cut-off frequency of the filter. The difference ΔV of the D.C. voltages produced by the two opponents is fed to the input of an automatic recording potentiometer (type PR 2000) [4]. The servo drive of the latter turns a horizontal arm about a fixed vertical spindle, the free end of the arm moving towards the competitor producing the high-

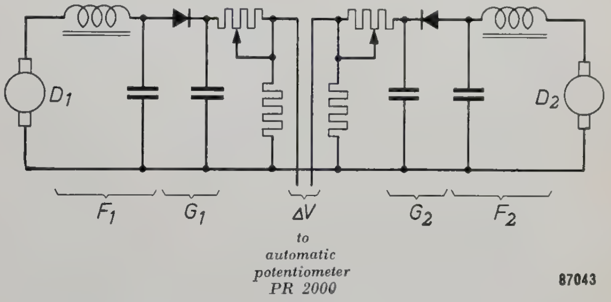

Fig. 5. Circuit diagram of the "Knijp van Jut" (Test your Clench). D_1, D_2 are hand-dynamos; F_1, F_2 low-pass filters and G_1, G_2 rectifiers.

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[3] E. A. van IJzeren, The dynamo pocket torch, Philips tech. Rev. **8**, 225-230, 1946.

[4] H. J. Roosdorp, An automatic recording potentiometer for industrial use. Philips tech. Rev. **15**, 189-220, 1953/1954.

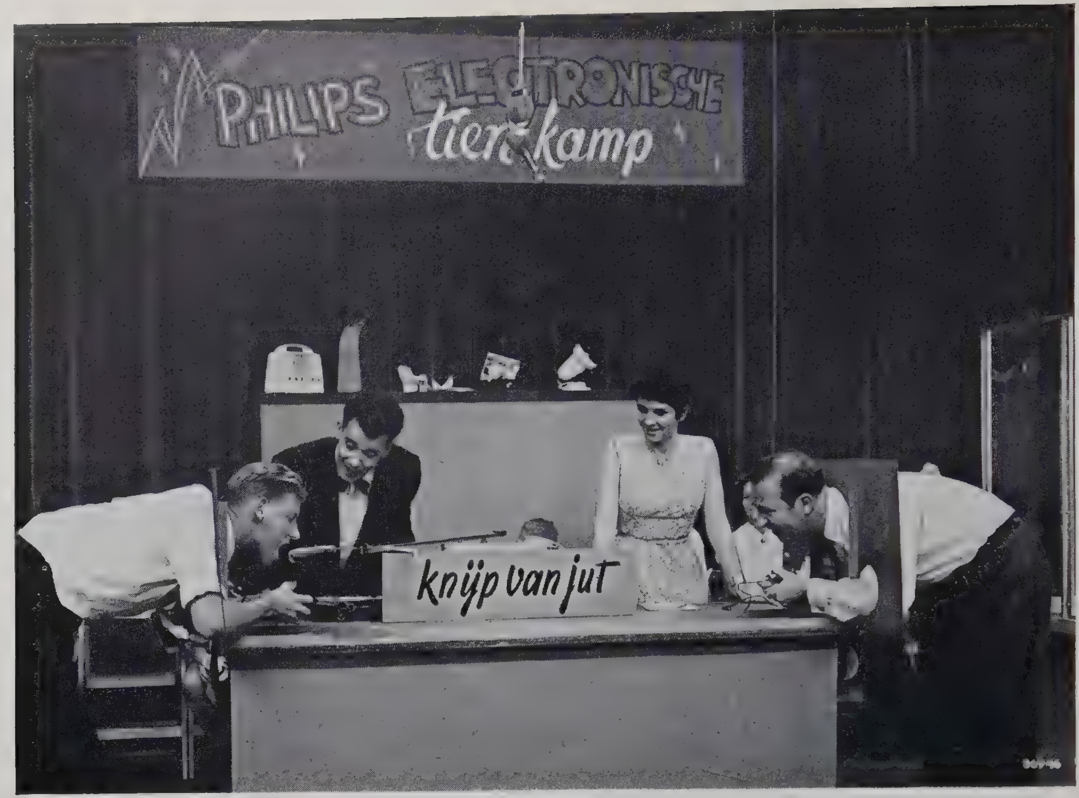

Fig. 6. "Knijp van Jut" (Test your Clench). This photograph was taken not at the scientific fair but during a TV transmission of the Philips "Electronische Tienkamp" ("Electronic Decathlon") at the E 55 exhibition held in Rotterdam in 1955.

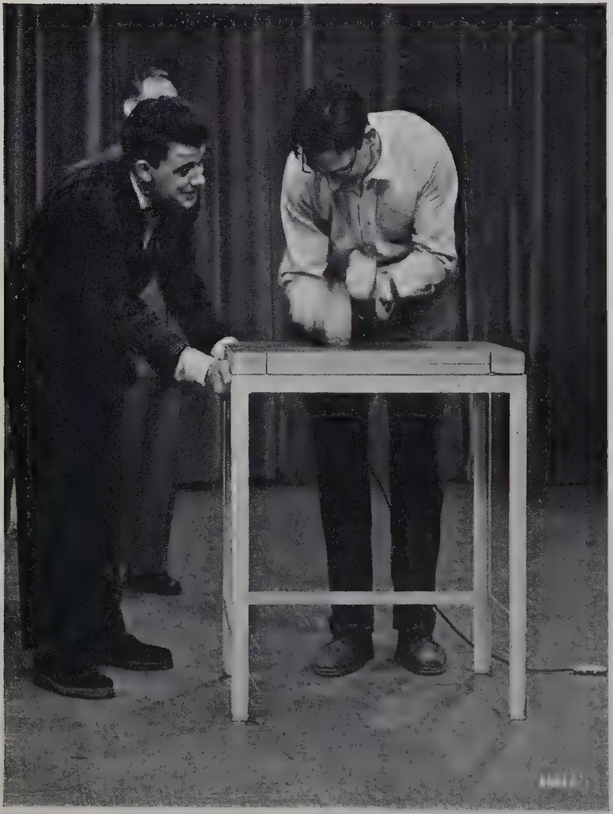

Fig. 7. The "Senilograph" (photograph taken at the E 55).

est voltage. If one of them succeeds in turning the arm completely round to his side a lollipop fastened on the arm lands in his mouth and he is the winner (*fig. 6*).

Senilograph. This competition involves beating with the fists on a table, under which is fitted an electronic counting meter which records the number of sufficiently strong blows within a stipulated length of time (*fig. 7*).

Phililoei — a contest in screaming power. Two amplifiers (see *figs. 8* and *9*) amplify the signals from two microphones each of which is held by one of the contestants. The A.C. voltages supplied by the amplifiers are rectified and each charges a condenser via a resistor whereby the voltage of the condenser gradually increases. When the condenser, for example C_1, reaches a potential of 125 V, a neon lamp N_1 connected across it ignites and the condenser begins to discharge via a relay in series with N_1. This closes contact K_1 which causes C_1 to discharge rapidly through the coil Rel_1 of a multi-contact switch W which thereby changes one position. When the voltage of C_1 drops below the working voltage of the neon lamp, K_1 re-opens and C_1 begins to re-charge. The multi-contact switch has two opposed coils corresponding to the two microphones, which cause the switch to rotate in opposite directions. A row of lamps L signals the position of the switch, giving visible indication of the louder and more persistent scream. It turned out that the greatest difficulty in this competition was that competitors had to retain sufficient self-control not to burst out laughing and so become short of breath.

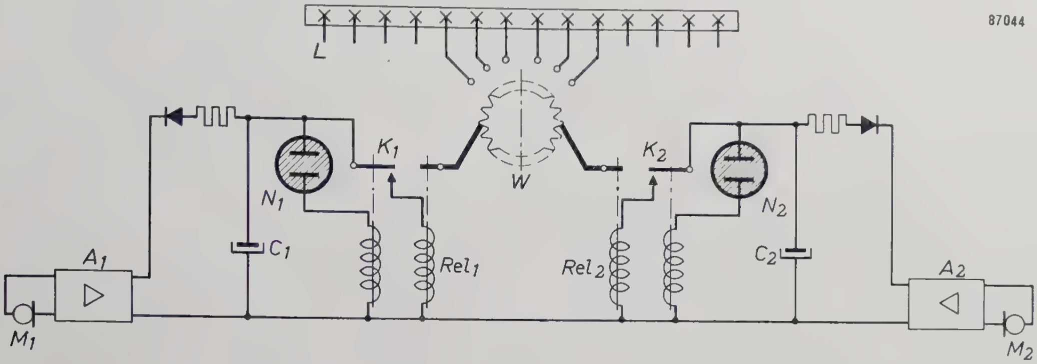

Fig. 8. Circuit of the "Phililoei". M_1, M_2 are microphones; A_1, A_2 amplifiers; C_1, C_2 condensers; N_1, N_2 neon lamps 85 A 2; K_1, K_2 relay contacts; Rel_1, Rel_2 exciting coils of the multi-contact switch W and L indicator lamps.

Magnoball — magnetic table football. The "ball", in the form of a vertically magnetized disk of ferroxdure [5]) (diameter 1″, height ½″), lies between two horizontal sheets of glass (about 30″×15″) with just sufficient clearance to slide between them.

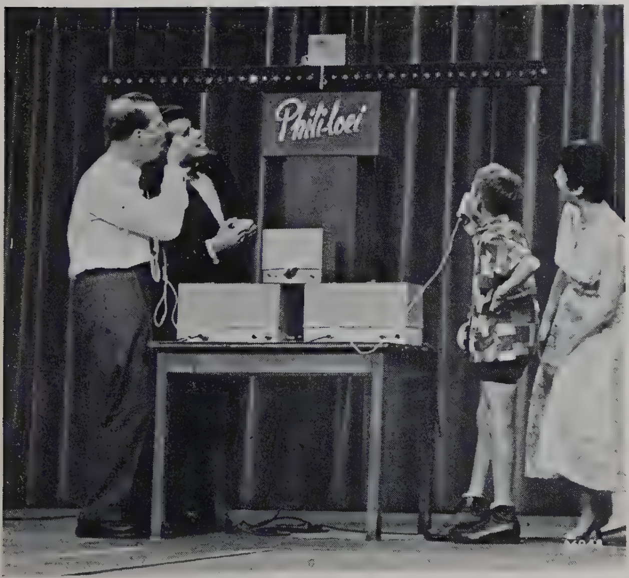

Fig. 9. "Phililoei" (photograph taken at the E 55). The two opponents each have a microphone in their hand. Above their heads is a row of indicator lamps.

The two players have each a disk in their hand identical to the "ball". One player holds his disk underneath the lower sheet whilst the other holds his on the upper sheet, each in such a way that both disks exert a repelling force on the "ball" (*fig. 10*). The field of play is bounded by a frame of ferroxdure, magnetized vertically in the same sense as the ball: consequently the ball never rests against the wall but is always repelled back into the field. The ferroxdure frame is absent at the two goals and each player tries to steer the ball into his opponent's goal. This game is surprisingly fast and requires great agility because of the somewhat unpredictable movements of the ball due to the action at a distance. Goals are shown electrically.

All the above mentioned attractions are of a competitive nature. We may now mention a few attractions which were organized just for entertainment.

Nilando the calculating genius. The "genius" was in

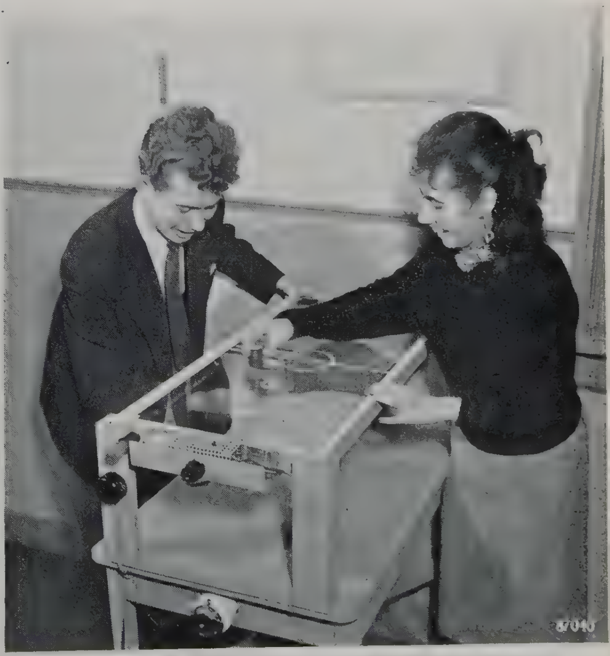

Fig. 10. "Magnoball", a magnetic table football game for two persons.

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[5]) Also known as Magnadur. See Philips tech. Rev. **13**, 194-208, 1951/1952.

touch with a computing room by means of an induction loop [6]) and a concealed hearing aid. Consequently he was able to solve the most difficult problems in the twinkling of an eye.

Boffin mirror. An electronic distorting mirror in which people saw their own image, picked up by a TV camera en reproduced with adjustable distortion on a TV screen.

Lotty — the right type. Lotty was a tele-typewriter used in the laboratory for printing the output from an electronic computer. At the fair the orders were fed into it in the form of punched teleprinter tapes. The tapes were prepared beforehand and contained answers to various questions which could be put to Lotty. The tape containing the answer was fed into the instrument by a concealed accomplice, so that Lotty typed out the correct answer. By suitable arrangements of characters and spaces she was able to produce caricatures of well-known personalities (*fig. 11*).

As may be gathered from the photographs shown here, the attractions of the fair also did service on a subsequent occasion, namely at the Dutch exhibition "Energy, 1955" held in Rotterdam last year. A number of the competitive sideshows were made available to the Publicity Department of Philips for an experimental commercial TV programme which ran successfully for several months at the "E 55". The Philips programme, entitled the "Electronic Decathlon", gave members of the audience the opportunity to test their skill and strength on these games. The programme ran daily from 15 July to 3 September 1955, and was an unqualified success. The popularity of these electronic playthings show not only their fascination for the contestants but also their considerable entertainment value for audiences.

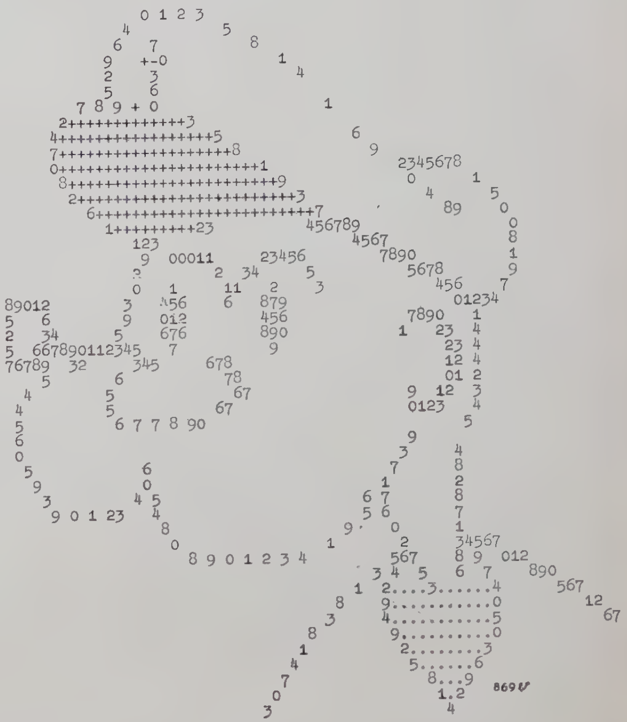

Fig. 11. A well-known figure caricatured by Lotty.

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[6] See p. 42 of P. Blom, An electronic hearing-aid, Philips tech. Rev. 15, 37-48, 1953/1954.

ABSTRACTS OF RECENT SCIENTIFIC PUPLICATIONS BY THE STAFF OF N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprints of these papers not marked with an asterisk * can be obtained free of charge upon application to Philips Electrical Ltd., Century House, Shaftesbury Avenue, London W.C. 2.

2246: J. F. Klinkhamer: Lineaire vierpolen met voorgeschreven overdrachtsimpedantie (De Ingenieur **67**, E1-E8, 1955). (Linear four-poles with specified transfer impedance functions; in Dutch).

After a discussion of the circumstances in which the synthesis of a four-pole with a prescribed transfer impedance is of practical importance, the synthesis is treated for the case that a) the prescribed transfer impedance is given as a rational function of frequency; b) the frequencies of infinite attenuation are real frequencies; c) the four-pole is allowed to consist of a reactive network together with one terminating resistance. After this a short discussion is given on the problem of deriving a feasible rational transfer impedance function with a prescribed attenuation curve, within given tolerances.

2247: G. W. van Santen: Het automatisch regelen van de temperatuur in een melkpasteur met behulp van electronische apparatuur (De Ingenieur **67**, 051-055, 1955, No. 14) (Automatic control of temperature in a milk pasteurizing plant by means of electronic equipment; in Dutch).

Several systems exist for pasteurizing milk; the best one, using a temperature of 72 °C, requires a very close temperature control. An installation with electronic controlling instruments is described; the practical difficulties arising from various disturbances are mentioned. Special attention is paid to disturbances caused by the action of the flow diversion valve. The electronics behind the Philips controlling and recording equipment are outlined.

2248: J. M. Stevels, C. van Amerongen and J. Volger: Dielectric losses in quartz at low temperatures (Z. physik. Chem. **3**, 382-385, 1955, No. 5-6).

Investigations of the deformation losses in quartz at temperatures of about 50 °K and at frequencies of a few kc/s. Clear crystals show several relaxation effects, obviously correlated with lattice defects of different types, the relaxation time being determined by an activation energy of the order of 0.1 eV. Quartz coloured by irradiation in the laboratory, on the other hand, and also natural smoky quartz, shows relatively high dielectric losses at very low temperatures particularly below 20 °K. Amethyst, however, differs essentially from the other crystals investigated.

2249: H. G. van Bueren and P. Jongenburger: Resistivity changes by plastic deformation of polycrystalline metals (Nature **175**, 544-545, 1955, No. 4456).

Annealed polycrystalline wires of copper and silver were plastically deformed at 20 °K and 77 °K by first extending them to about 10%, then applying a twist corresponding to a surface shear of about 0,1, and subsequently extending them further. The increase in resistivity caused by this deformation was measured. It is caused by the formation of point-defects (vacancies, interstitial atoms) and line-defects (dislocations). The observations, combined with recent theories, lead to the conclusion that the major part of the resistivity increase caused by plastic deformation is contributed by the point-defects.

2250: P. Jongenburger: Some remarks on the extra-resistivity due to interstitial atoms in copper (Nature **175**, 545-546, 1955, No. 4456).

Previous calculations of the extra-resistivity due to the presence of vacancies in copper are extended to the case of interstitial atoms. The displacements of the neighbouring atoms are very large in this case, and a rough estimation shows them to be responsible for a large part of the extra-resistivity. The contribution of the interstitial atom itself is also calculated. Supposing both parts of the extra-resistivity to be additive the conclusion is drawn that interstitial atoms cause an increase in resistivity of about 5 $\mu\Omega$ cm per atom percent.

2251: R. Vermeulen: Validity of hypotheses (Synthese **9**, 385-394, 1955).

Remarks on the evaluation of the validity of hypotheses by means of Bayes' theorem. The author develops an idea of H. Jeffreys and gives an example to show that the results agree with common sense.

2252: D. Kleis: Experimente zur Verbesserung des Raumwirkung von Schall (Elektronische Rundsch. **9**, 64-68, 1955, No. 2). (Experiments to improve the acoustical qualities of halls; in German).

An account of the work described in Philips tech. Rev. **17**, 258-266, 1955/56, No. 9.

2253: M. C. Teves: Image intensification: instrumental aspects (Brit. J. Radiology **28**, 216-217, April 1955).

Short description of the image intensifier tube as now produced by Philips, a development based on requirements laid down by Chamberlain in Brit. J. Radiology, 1942. The factors limiting the contrast detail rendering and a fundamental factor limiting the minimum X-ray intensity are discussed briefly. Finally the use and advantages of the tube for photographic and cinematographic recording are discussed. See also Philips tech. Rev. **17**, 69-77, 1955/56 (No. 3).

2254: J. Feddema: Image intensification: some possible diagnostic applications in cineradiography (Brit. J. Radiography **28**, 217-220, 1955/56, No. 3).

Clinical experiences in the use of the X-ray image intensifier for diagnostic purposes and comparison with other X-ray diagnostic techniques. See also Philips tech. Rev. **17**, 88-93, 1955 (No. 3).

2255: A. van Weel: A direct-indicating phase meter (J. Brit. Instn. Rad. Engrs. **15**, 143-152, 1955, No. 3).

A phase-measuring method is described, in which the unknown phase angle is introduced in an oscillating circuit, thus influencing the oscillator frequency. The phase angle can be determined by measuring the frequency shift of the oscillator. Both very small (10^{-4} radian) and larger phase angles (up to 2π radians) can be measured in this way. A description of a complete phase meter is given. See also Philips tech. Rev. **15**, 307-316, 1953/54.

2256: H. J. G. Meyer: On the theory of radiative transitions of trapped electrons in polar crystals (Physica **21**, 253-268, 1955, No. 4).

Further consideration of the theory of the properties of F-centre absorption bands in ionic crystals (see previous article by the same author, these abstracts No. **2223**).

2257: H. O. Huisman and A. Smit: A new synthesis of 2,4,5-trichlorophenoxyacetic acid (2,4,5-T) (Rec. Trav. chim. Pays-Bas **74**, 155-160, 1955, No. 2).

Description of a new synthesis for the above-named compound. A mixture of the insecticidally inactive α and β isomers of hexachlorocyclohexane, obtained as a by-product in the manufacture of the γ isomer is converted into 1,2,4-trichlorobenzene. After converting the 1,2,4-trichlorobenzene into 2,5-dichlorophenol, the latter is reacted with monochloroacetic acid to 2,5-dichlorophenoxyacetic acid and chlorinated into 2,4,5-T. The overall yield of 2,4,5-T calculated on hexachlorocyclohexane is about 40-50 mole %.

2258: N. W. H. Addink: An improved method for the spectrochemical determination of zinc in blood (Rec. Trav. chim. Pays-Bas **74**, 197-205, 1955, No. 3).

A method for determining the concentration of zinc in blood with greater precision than the earlier methods described by the same author (see these abstracts, Nos. **1947** and **1972**). The main points of the method are: a) complete evaporation of the metal out of a weighed amount (10 mg) of blood ash; b) a constant arc temperature independent of a varying sodium content of the blood ash; c) a rotating anode of a definite shape. By using this method, concentration values have been determined with a standard deviation of 5%. As for the absolute concentration of zinc in various blood samples, values found by applying the method of successive additions, agreed satisfactorily with those determined chemically.

2259: P. B. Braun and J. H. N. van Vucht: Al-Th intermetallic compounds, I (Acta Cryst. **8**, 117,1955, No. 2).

The system thorium-aluminium has been investigated. Six intermetallic compounds have been found. In this paper the structure and lattice constants are given of Al_3Th (hexagonal), Al_2Th (hexagonal) and Al_2Th_3 (tetragonal).

2260: P. B. Braun and J. H. N. van Vucht: Al-Th intermetallic compounds, I (Acta Cryst. **8**, 246, 1955, No. 4).

The structure and lattice constants are given of the compounds Al-Th (orthorhombic) and $AlTh_2$ (tetragonal) and a probable unit cell for a compound with the probable composition Th_4Al_7 which is stable only in a narrow temperature range at about 1300 °C.

2261: J. B. de Boer: A "Duplo" headlight with asymmetric passing beam (Light and Lighting **48**, 137-141, 1955, No. 4).

Description of a car headlamp bulb (briefly described in Philips tech. Rev. **16**, 351-352, 1954/1955, No. 12) in which the cap under the "dip" filament is partially cut away. In conjunction with a suitable headlight glass, this bulb largely combines the advantages of the European and the American dipping systems. Tests with the lamps are described.

2262: J. P. Spruyt: Accelerated stability test for vitamin A in oils and fats by means of surface enlarging at room temperature (J. Amer. Oil Soc. **32**, 197-200, 1955, No. 5).

Abbreviated version of the article abstracted below: see No. **2263**.

2263: J. P. Spruyt: Versnelde houdbaarheidsproef door middel van oppervlaktevergroting bij kamertemperatuur voor vitamine A in oliën en tranen (Voeding **16**, 416-429, 1955, No. 5). (Title as in **2262**; in Dutch).

A new, simple and rapid method for determining the stability of vitamin A in oils and fats. It is performed at a constant temperature of 20 °C; the acceleration is obtained by enlarging the surface exposed to the air by spreading the oil over a neutral carrier.

2264: J. L. Meijering: Het systeem koper-nikkel-chroom (Chem. Weekblad **51**, 438-441, 1955, No. 23) (The system copper-nickel-chromium; in Dutch).

In the ternary system Cu-Ni-Cr, two face-centered cubic phases appear, besides the body-centered phase rich in Cr (see also J. Inst. Metals **84**, 118-120, 1955/1956). This looks at first sight rather unexpected, but is shown to be in line with the result of thermodynamical calculations based on the regular solution model.

2265: H. G. van Bueren: Elektrischer Widerstand und plastische Deformation von Metallen (Z. Metallk. **46**, 272-282, 1955, No. 4). (Electrical resistance and plastic deformation of metals; in German).

A review is given of the observations concerning the relation between electrical resistivity and plastic flow at low temperatures of pure metals, especially copper. A theoretical foundation of the observed relationship is proposed. There is strong evidence that the resistivity increase is caused in the first instance by the vacancies and interstitial atoms formed in the lattice during plastic flow. The recovery phenomena of the electrical conductivity on heating are described and compared to similar phenomena in irradiated and quenched metals. Five recovery steps at different temperatures are indicated. A tentative explanation is proposed in terms of various diffusion phenomena of the above-mentioned lattice defects.

2266: A. Claassen and A. Daamen: The photometric determination of cobalt by extraction with β-nitroso-β-naphthol (Anal. chim. Acta **12**, 547-553, 1955, No. 6).

An improved procedure is given for the photometric determination of cobalt by extraction of the α-nitroso-α-naphthol complex with chloroform. In contrast to former procedures it can be applied in the presence of large amounts of foreign elements, in particular iron and nickel. Interfering elements are gold, palladium and more than 25 mg of copper. The method has been applied with good results to the determination of cobalt in steels, non-ferrous alloys and in nickel.

2267: J. A. Lely: Darstellung von Einkristallen von Siliciumkarbid und Beherrschung von Art und Menge der im Gitter eingebauten Verunreinigungen (IUPAC, Colloq. Münster, Westf. 2-6 Sept. 1954, published by Chemie, Weinheim 1955) (Preparation of single crystals of silicon carbide and control of type and quantity of lattice impurities; in German).

Single crystals of the hexagonal form of silicon carbide (SiC) have been prepared by sublimation of commercial SiC in a protective gas at atmospheric pressure in a graphite furnace at a temperature of 2500-2600 °C (4500-4700 °F).

The sublimation of SiC is complicated by the fact that together with the molecular sublimation there occurs also dissociation into Si-vapour and solid graphite. By lining a graphite crucible on the inside with blocks of SiC one obtains a chamber in which clear, graphite-free SiC crystals are deposited. In 6-8 hours crystals are formed up to 10 mm wide and 2-3 mm thick. Very pure crystals ($\leq 10^{-5}\% = 10^{10}$ impurity atoms/cm^3) are practically colourless; when Al or B is built into the lattice they are blue and show hole (p-type) electrical conductivity. By introducing nitrogen or phosphorus the crystals acquire a green colour and show electron (n-type) conductivity.

The impurity to be introduced is added preferably to the protective gas (in most cases Argon) in the form of the element (N_2) or a volatile compound ($AlCl_3$, BCl_3, etc.). In this manner the crystals show a homogeneous impurity content. By changing over during a run from N_2 to $AlCl_3$, it is possible to obtain crystals with differently coloured zones, so that the blue zone shows p-type conductivity and the green zone n-type conductivity.

2268: Y. Haven: Concentration and association of lattice defects in NaCl (Rep. Conf. Defects in Crystalline Solids, pp. 261-272, Bristol 1954).

In NaCl containing small amounts of divalent impurities like Ca^{2+}, Cd^{2+}, Mn^{2+} etc., the (—) Na vacancies are captured by the (+) divalent ions. The concentration of associates (Ca^{2+}-Na vacancy) can be deduced rather accurately from dielectric measurements. The concentration of free ions can be deduced much less accurately from conductivity measurements. It appears that not only the equilibrium Ca^{2+} + Na vacancy $\rightleftarrows$ (Ca^{2+}-Na vacancy) plays a role, but also the associates consisting of more individuals. The different behaviour of Cd^{2+} and Ca^{2+} is largely determined by the segregation of mixed crystals for Cd^{2+}, rather than by the pair association. At room temperature the vacancies have still a relatively large mobility. The mobility of the associated vacancies seems to be higher than of free vacancies. The Cl^- vacancies seem to have a mobility about 6 times smaller than that of Na vacancies. Isotypic mixed crystals like NaCl-NaBr do not exhibit an excessive concentration of lattice defects.

R 274: A. Bril and H. A. Klasens: Phosphors for tricolour television tubes (Philips Res. Rep. **10**, 305-318, 1955, No. 5).

The efficiencies and spectral distributions were measured of various blue, green and red phosphors which might be considered for use as primaries in tricolour tubes. It is shown that it is mainly the red phosphor that determines the brightness of the screen and that a compromise has to be found between colour reproduction and light output. Of the known red phosphors, a silver-activated zinc-cadmium sulphide and a copper-activated zinc-cadmium selenide can be expected to produce higher efficiencies for near white colours than is possible with the commonly used manganese-activated zinc phosphate.

R 275: J. van den Boomgaard: Zone-melting processes under influence of the atmosphere (Philips Res. Rep. **10**, 319-336, 1955, No. 5).

The mathematical description of the zone-melting process, as given by Pfann and by Reiss concerns only the cases of ingots consisting of an element or a non-decomposing compound with non-volatile solutes. In this paper the theory is extended to elements containing volatile solutes. It is shown that it is possible to bring a volatile impurity element homogeneously into an ingot by means of zone melting under a constant vapour pressure of that element, providel that diffusion in the liquid phase is rapid.

R 276: B. H. Schulz: Analysis of the decay of photoconductance in germanium (Philips Res. Rep. **10**, 337-348, 1955, No. 5).

A method of analysing surface and volume effects in the recombination of injected charge carriers is described. The influence of the capacity of a surface double layer is discussed and some of the results of measurements are given.

R 277: A. G. Th. Becking, H. Groendijk and K. S. Knol: The noise factor of four-terminal networks (Philips Res. Rep. **10**, 349-357, 1955, No. 5).

In this paper a simple formula is given for calculating the noise factor of a linear four-terminal network with internal noise sources for any input circuit and any value of signal source impedance. This formula is derived from the theorem that the noise behaviour of the network can be characterized by two noise sources introduced at the input of the network. A proof of this theorem is given. The noise sources and their mutual correlation can be denoted by four quantities, which may be determined by four simple measurements.

R278: H. K. Hardy and J. L. Meijering: Phase separation in quaternary regular solutions (Philips Res. Rep. **10**, 358-400, 1955, No. 5).

Calculation of the phase separation of quaternary solutions based on the "regular" approximation for the free energy (see earlier calculations on ternary solutions: these abstracts **R149** and **R166**). The regular approximation includes the Gibbs expression for the entropy of mixing and six terms representing the heat of mixing. Each of these six terms consists of the product of one interaction parameter (corresponding to one of the six binary boundary systems) and the concentrations of the two components concerned in that binary system. In particular, the existence of quaternary critical points and the shift of the critical curves with temperature are investigated.

R 279: G. Diemer, H. A. Klasens and J. G. van Santen: Solid state image intensifiers (Philips Res. Rep. **10**, 401-424, 1955, No. 6).

A theory is given of the characteristics of solid-state image-intensifying screens consisting of a photoconducting and an electroluminescent layer, including the influence of positive internal feedback and negative electrical feedback. The possibilities of brightness as well as contrast amplification are discussed. With intermittent irradiation it is possible to increase the amplification factor by storage, due to the decay, or by triggering of a feedback amplifying-screen. In the latter case the gradation need not be lost. The dimensioning of the parameters with regard to specific applications (e.g. radar, X-ray images, television) is discussed. Preliminary experimental results are in agreement with theory. With X-rays a brightness-amplification factor of 30 was obtained with respect to a normal fluorescent X-ray screen.

R 280: Th. A. J. Payens: Ionized monolayers (Philips Res. Rep. **10**, 425-481, 1955, No. 6).

This thesis deals with monomolecular layers of long-chain weak electrolytes at oil-water and air-water interfaces. The ionization of such films sets up an electronic double layer at the interface. The contribution of this double layer to the interfacial tension is derived from general thermodynamical considerations. The theory is put to the test with monolayers of various fatty acids, stearyl phosphonic acid and octylamine on sub-phases of various *p*H and salt concentration. Deviations from the theory at salt concentrations exceeding 0.1 N NaCl can be explained by the penetration of the counter-ions between the film charges.